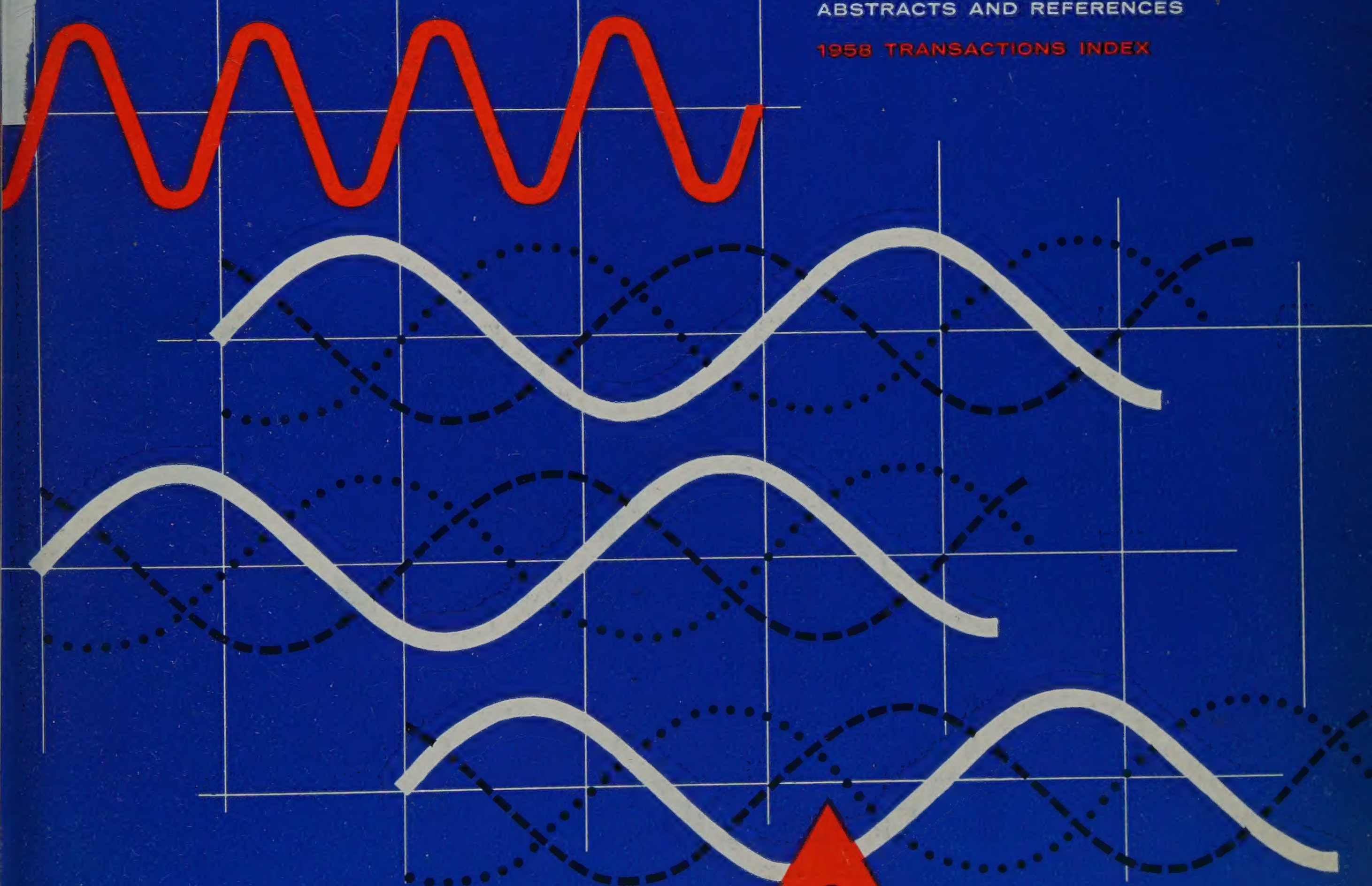
april 1959

the institute of radio engineers

Proceedings of the IRE

in this issue



A New Computing Tool—
Multi-Phase Subharmonics: Page 516

UTC NEW EXPANDED DO-T AND DI-T SERIES

Revolutionary transistor transformers hermetically sealed to MIL-T-27A Specifications.

UTC DO-T and DI-T transistor transformers provide unprecedented power handling capacity and reliability coupled with extremely small size. Comparative performance with other available products of similar size are shown in the curves (based on setting output power at 1 KC, then maintaining same input level over frequency range). The new expanded series of units cover virtually every transistor application.

DO-T

ACTUAL SIZE

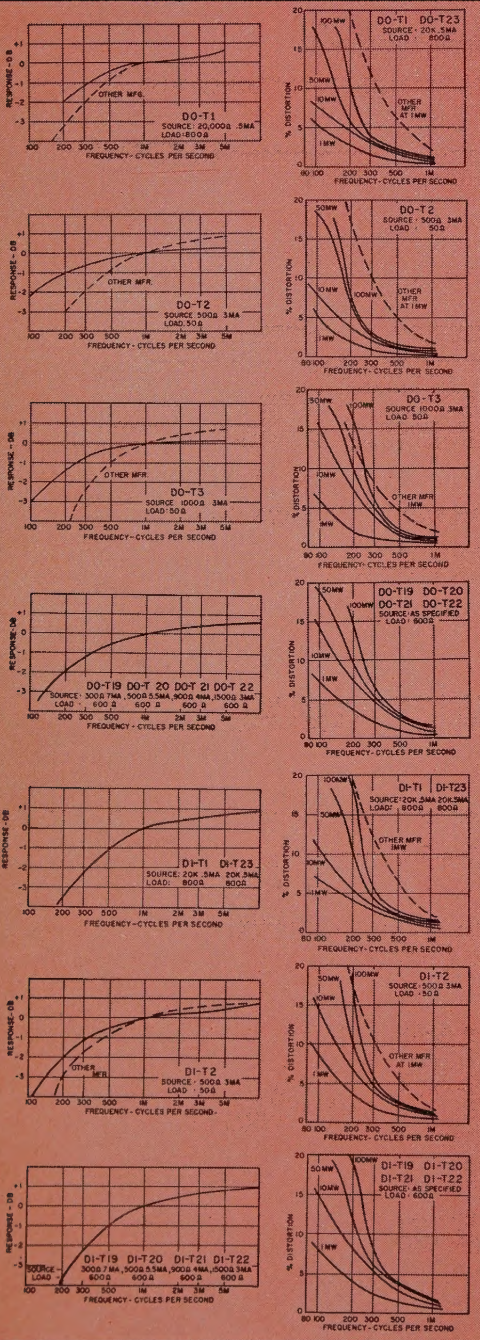
5/16 Dia. x 13/32, 1/10 Oz.

High Power Rating . . . up to 100 times greater.
Excellent Response . . . twice as good at low end.
Low Distortion . . . reduced 80%.
High Efficiency . . . up to 30% better.
Moisture Proof . . . hermetically sealed to MIL-T-27A.
Rugged . . . completely cased.
Anchored Leads . . . withstand 10 pound pull test.
Printed Circuit Use...plastic insulated leads.

DI-T

ACTUAL SIZE

5/16 Dia. x 1/4, 1/20 O



DO-T No.	MIL Type	Application	Pri. Imp.	D.C. Ma.‡ in Pri.	Sec. Imp.	Pri. Res. DO-T	Pri. Res. DI-T	Level Mw.	DI-T No.
DO-T1	TF4RX13YY	Interstage	20,000 30,000	.5 .5	800 1200	850	815	50	DI-T1
DO-T2	TF4RX17YY	Output	500 600	3 3	50 60	60	65	100	DI-T2
DO-T3	TF4RX13YY	Output	1000 1200	3 3	50 60	115	110	100	DI-T3
DO-T4	TF4RX17YY	Output	600	3	3.2	60		100	
DO-T5	TF4RX13YY	Output	1200	2	3.2	115	110	100	DI-T5
DO-T6	TF4RX13YY	Output	10,000	1	3.2	1000		100	
DO-T7	TF4RX16YY	Input	200,000	0	1000	8500		25	
DO-T8	TF4RX20YY	Reactor 3.5 Hys. @ 2 Ma. DC, 1 Hy. @ 5 Ma. DC				630			
	TF4RX20YY	Reactor 2.5 Hys. @ 2 Ma. DC, .9 Hy. @ 4 Ma. DC				630			DI-T8
DO-T9	TF4RX13YY	Output or driver	10,000 12,500	1 1	500 CT 600 CT	800	870	100	DI-T9
DO-T10	TF4RX13YY	Driver	10,000 12,500	1 1	1200 CT 1500 CT	800	870	100	DI-T10
DO-T11	TF4RX13YY	Driver	10,000 12,000	1 1	2000 CT 2500 CT	800	870	100	DI-T11
DO-T12	TF4RX17YY	Single or PP output	150 CT 200 CT	10 10	12 16	11		500	
DO-T13	TF4RX17YY	Single or PP output	300 CT 400 CT	7 7	12 16	20		500	
DO-T14	TF4RX17YY	Single or PP output	600 CT 800 CT	5 5	12 16	43		500	
DO-T15	TF4RX17YY	Single or PP output	800 CT 1070 CT	4 4	12 16	51		500	
DO-T16	TF4RX13YY	Single or PP output	1000 CT 1330 CT	3.5 3.5	12 16	71		500	
DO-T17	TF4RX13YY	Single or PP output	1500 CT 2000 CT	3 3	12 16	108		500	
DO-T18	TF4RX13YY	Single or PP output	7500 CT 10,000 CT	1 1	12 16	505		500	
DO-T19	TF4RX17YY	Output to line	300 CT	7	600	19	20	500	DI-T19
DO-T20	TF4RX17YY	Output or line to line	500 CT	5.5	600	31	32	500	DI-T20
DO-T21	TF4RX17YY	Output to line	900 CT	4	600	53	53	500	DI-T21
DO-T22	TF4RX13YY	Output to line	1500 CT	3	600	86	87	500	DI-T22
DO-T23	TF4RX13YY	Interstage	20,000 CT 30,000 CT	.5 .5	800 CT 1200 CT	850	815	100	DI-T23
DO-T24	TF4RX16YY	Input (usable for chopper service)	200,000 CT	0	1000 CT	8500		25	
DO-T25	TF4RX13YY	Interstage	10,000 CT 12,000 CT	1 1	1500 CT 1800 CT	800	870	100	DI-T25
DO-T26	TF4RX20YY	Reactor 6 Hy. @ 2 Ma. DC, 1.5 Hy. @ 5 Ma. DC				2100			
	TF4RX20YY	Reactor 4.5 Hy. @ 2 Ma. DC, 1.2 Hy. @ 4 Ma. DC				2300			DI-T26
DO-T27	TF4RX20YY	Reactor 1.25 Hy. @ 2 Ma. DC, .5 Hy. @ 11 Ma. DC				100			
	TF4RX20YY	Reactor .9 Hy. @ 2 Ma. DC, .5 Hy. @ 6 Ma. DC				105			DI-T27
DO-T28	TF4RX20YY	Reactor .3 Hy. @ 4 Ma. DC, .15 Hy. @ 20 Ma. DC				25			
	TF4RX20YY	Reactor .1 Hy. @ 4 Ma. DC, .08 Hy. @ 10 Ma. DC				25			DI-T28
DO-T29	TF4RX17YY	Single or PP output	120 CT 150 CT	10 10	3.2 4	10		500	
DO-T30	TF4RX17YY	Single or PP output	320 CT 400 CT	7 7	3.2 4	20		500	
DO-T31	TF4RX17YY	Single or PP output	640 CT 800 CT	5 5	3.2 4	43		500	
DO-T32	TF4RX17YY	Single or PP output	800 CT 1,000 CT	4 4	3.2 4	51		500	
DO-T33	TF4RX13YY	Single or PP output	1,060 CT 1,330 CT	3.5 3.5	3.2 4	71		500	
DO-T34	TF4RX13YY	Single or PP output	1,600 CT 2,000 CT	3 3	3.2 4	109		500	
DO-T35	TF4RX13YY	Single or PP output	8,000 CT 10,000 CT	1 1	3.2 4	505		500	
DO-T36	TF4RX13YY	Isol. or Interstage	10,000 CT	1	10000 CT	950	970	500	DI-T36

DO-TSH Drawn Hipermalloy shield and cover for DO-T's, provides 25 to 30 db shielding, for DI-T's **DI-TSH**
‡DCMA shown is for single ended useage (under 5% distortion—100MW—1KC) . . . for push pull, DCMA can be any balanced value taken by .5W transistors (under 5% distortion—500MW—1KC)
*DO-T units have been designed for transistor application only . . . not for vacuum tube service. Pats. Pend.

And Special Units to Your Specifications

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EXPORT DIVISION: 13 EAST 40th STREET, NEW YORK 16, N. Y. CABLES: "ARLAB"

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VOLUME 47, NUMBER 4

April, 1959

published monthly by The Institute of Radio Engineers, Inc.

Proceedings of the IRE ®

contents

published monthly by The Institute of Radio Engineers, Inc.

Proceedings of the IRE®

continued

COVER A novel computing scheme in which digital information is represented by the phase of a subharmonic wave is described on page 516 of this issue.

PROCEEDINGS OF THE IRE, published monthly by The Institute of Radio Engineers, Inc. at 1 East 79 Street, New York 21, N. Y. Manuscripts should be submitted in triplicate to the Editorial Department. Responsibility for contents of papers published rests upon the authors, and not the IRE or its members. All republication rights, including translations, are reserved by the IRE and granted only on request. Abstracting is permitted with mention of source.

Thirty days advance notice is required for change of address. Price per copy: members of the Institute of Radio Engineers, one additional copy $1.25; non-members $2.25. Yearly subscription price: to members $9.00, one additional subscription $13.50; to non-members in United States, Canada, and U. S. Possessions $18.00; to non-members in foreign countries $19.00. Second-class postage paid at Menasha, Wisconsin under the act of March 3, 1879. Acceptance for mailing at a special rate of postage is provided for in the act of February 28, 1925, embodied in Paragraph 4, Section 412, P. L. and R., authorized October 26, 1927. Printed in U.S.A.

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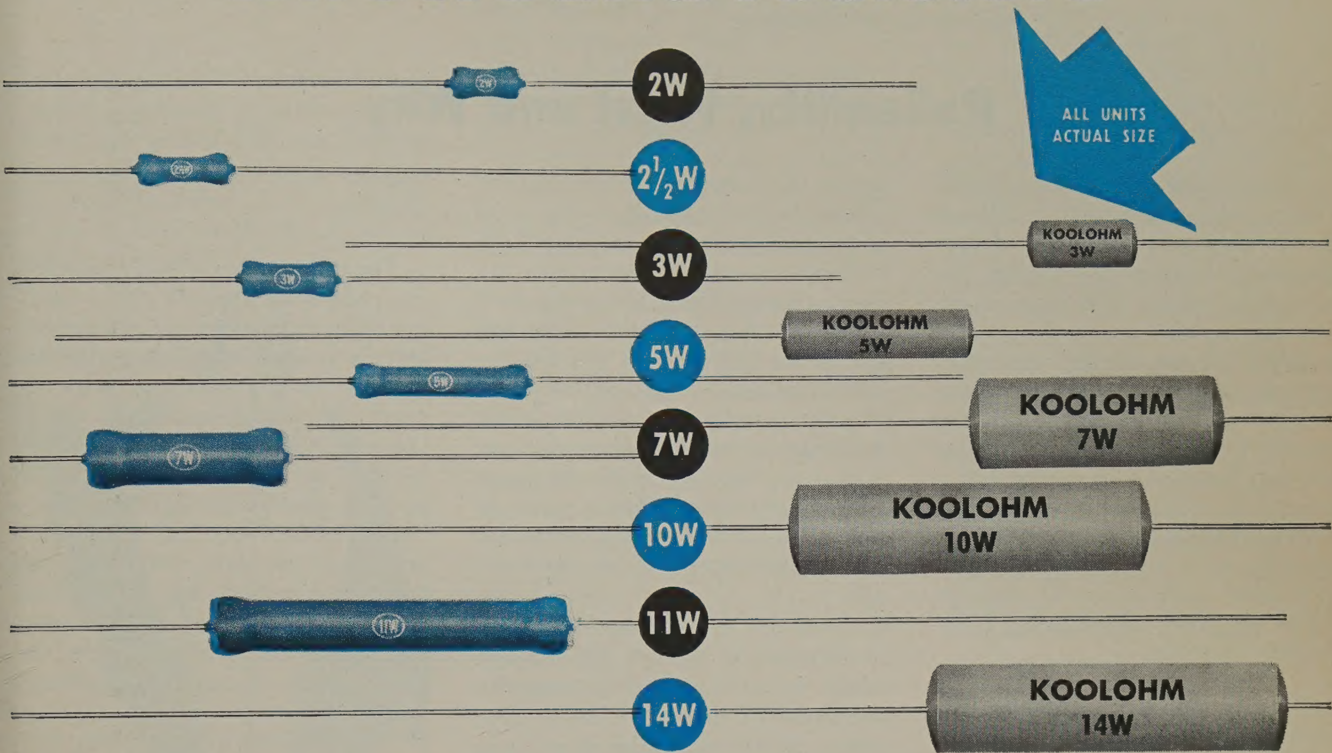
in these two dependable wirewound resistors



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Sprague's new improved construction gives even greater reliability and higher wattage ratings to famous Blue Jacket miniature axial lead resistors.

A look at the small *actual sizes* illustrated, emphasizes how ideal they are for use in miniature electronic equipment with either conventional wiring or printed wiring boards.

Get complete data on these dependable minified resistors, **write for Engineering Bulletin 7410.**

TAB-TYPE BLUE JACKETS: For industrial applications, a wide selection of wattage ratings from 5 to 218 watts are available in Sprague's famous Tab-Type Blue Jacket close-tolerance, power-type wirewound resistors. Ideal for use in radio transmitters, electronic and industrial equipment, etc. **For complete data, send for Engineering Bulletin 7400A.**

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New Koolohm construction features include welded leads and winding terminations—Ceron ceramic-insulated resistance wire, wound on special ceramic core—multi-layer non-inductive windings or high resistance value conventional windings—sealed, insulated, non-porous ceramic outer shells—*aged-on-load* to stabilize resistance value.

You can depend upon them to carry maximum rated load for any given physical size.

Send for Engineering Bulletin 7300 for complete technical data.

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THE MARK OF RELIABILITY

SPRAGUE COMPONENTS: RESISTORS • CAPACITORS • MAGNETIC COMPONENTS • TRANSISTORS
INTERFERENCE FILTERS • PULSE NETWORKS • HIGH TEMPERATURE MAGNET WIRE • PRINTED CIRCUITS

AIL'S Project STAR (Space Technology and Advanced Research) has had to face very serious reliability problems. We felt that some of the work performed on that project should appear as a part of our "advertisements" and asked Karle Packard, Manager of Reliability on AIL's Project STAR to prepare the next four articles.

These articles are not intended for the reliability experts but we hope that they will serve as a guide to the design engineer and alert him to the fundamentals and importance of reliability.

Reliability, What and Why

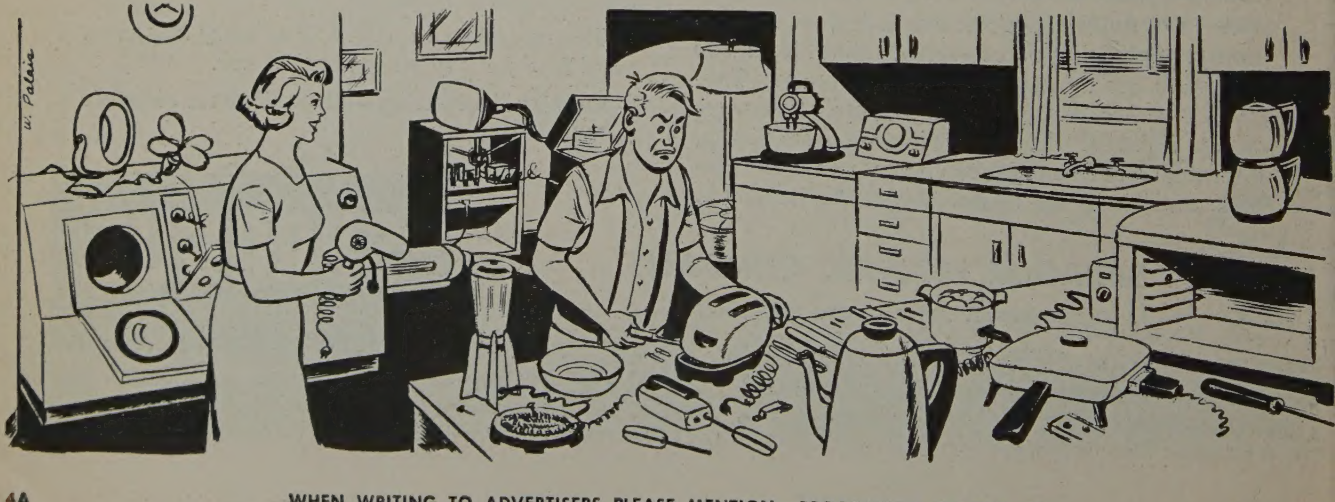
Suppose that you are an avid TV viewer and suppose also that TV sets broke down on the average of once for every 4 hours of viewing. Assume that it required 6 days, on the average, to repair the set. In order to watch TV every evening you would have to own 7 sets and pay a tremendous repair bill. It is certain that no one would tolerate this situation for very long—TV would be only a laboratory curiosity. It is precisely such a situation, however, which the military users of electronic equipment now face. It is this situation which the study of reliability is intended to remedy.

It is of interest to consider why there is suddenly a need for special consideration of the problem. For all but the most recent years of human history, the performance expected from man's implements was quite low and the life realized was "long," both because it just happened to be so in terms of man's lifetime and because he had no reason to expect anything else. Recent technological advances have been inextricably tied to more and come complex implements or devices. These, in general, have been a synthesis from simpler devices which had a satisfactory life. Any entity which, in order to function, requires all of its parts to function will always be less stable than any of its parts and, therefore, as devices become more complex they tend to become shorter lived. Advances have been at such a rapid rate that there has been relatively little time for improvement of the basic devices in order that complex equipment may last a sufficient time. Within the past few years we have come to the point where equipment failures occur so often that, in many cases, the utility of the equipment becomes questionable.

Although it has not yet reached really serious proportions, consider the problem of the ordinary consumer. With two or three dozen electrical appliances and mechanical contrivances in the average home, there are few times when all of them are in perfect operating condition. In this respect, the average home is seldom in perfect order. Many people spend a large part of their increased leisure time on these repairs and this situation is probably a major contributor to the "do-it-yourself" trend. In addition to the annoyance, U. S. consumers are spending (1957) \$16-billion a year for the repair and servicing of appliances. Although this creates work it is, nevertheless, wasteful.

The military electronic equipment situation is in a much worse condition. The maintenance of this equipment is costing us ten times the initial cost over a service life of five years. If civilian consumers had to "re-buy" their appliances every six months, our standard of living would be far below what it is. Our military "standard of living" cannot be compromised, however, and we are forced to support this tremendous waste. Each and every one of us must pay for it. The design engineer, interested only in new developments, must pay a much higher price than others. He must struggle along with inadequate funds because the maintenance of operational equipment must take precedence over new developments. The product designer who must cope with complaints from the field also pays a high price. This is just, however, as these are the people on whom the reliability of the equipment most depends. It is up to them to improve the situation.

An aspect more serious than the economic view, although not so immediately apparent, is the potential threat to security. We have become,



and will be more so in the future, highly dependent on complex electronic equipment for our defense. Despite this dependence we now have the situation where, at any one time, about 84 percent of our radars are down for repairs.

In "modern dress," with apologies to Benjamin Franklin,

A little neglect may breed mischief;
for want of a diode, a relay failed;
for want of the relay, a radar failed;
for want of a radar, a missile turned;
for want of the missile, a city burned.

These then are the two basic reasons for the emphasis on reliability in military electronic equipment; the high cost of maintenance and the undependability in operation.

Having examined the reliability problem in somewhat vague terms, although in terms of experience, it is now necessary to become more specific. Reliability, an aspect of engineering common to all fields of engineering, is complementary to the design aspect as interpreted in its narrow sense. Whereas the design aspect is primarily an attempt to create a device which will perform some desired function, the reliability aspect is an attempt to ensure that the device so designed will so perform under all expected conditions. Thus the subject of reliability involves a determination of the survival of the individual device under the expected conditions of environment, operation and maintenance. It may also involve a study of the survival of the species, or equipment type, under the conditions of fabrication and assembly. This latter type of study is known as reproducibility and is of concern only when reasonably large production quantities are contemplated.

A period of time during which a device operates satisfactorily may be terminated by any one of a large number of independent causes. For example, a vacuum tube has about 12 modes of failure, and an electronic instrument has a much larger number, since failure may be caused by failure of any one of a large number of parts. Such failures, furthermore, depend on a large variety of contingencies such as the materials and processing of the parts, the assembly procedure, and, usually, a highly variable environment. This combination of an extremely diverse many-to-one causal relationship and the existence of so many contingencies leads to an exceedingly random distribution of failures. For this reason, in order to quantitatively describe the reliability of a device, it is necessary to utilize the concept of probability. The following definition may be used:

***Reliability* is the probability that a device will, under specified conditions, perform as specified.**

Many published definitions explicitly call out time of operation as a specified condition. This has been avoided, purposely, as being unnecessary and misleading since time is often of absolutely no interest.

With regard to the present division between the design and reliability aspects of engineering, it is probably evidence of immaturity. It is only apparent in the field of electronic engineering which is young and growing vigorously. In the older fields of civil and mechanical engineering there is not such a separation. In fact, in the mechanical engineering field, although there is no strict division, the engineer is chiefly concerned with reliability, leaving most "design work" to designers. Even in electronic engineering the division is often obscure and is at best a makeshift expedient. It has been necessitated by the rapid progress in this field which, by concentrating on design for greater performance, has prevented adequate consideration of the continuing life of an individual equipment. This has resulted in electronic equipment which has proven unsatisfactory in use and, though users' dissatisfaction, has forced a special consideration of the reliability of electronic equipment. It should be obvious to all, however, that a *reliable* device must be *designed* that way, that the engineer who originates it is in the best position to make it reliable, and that it isn't really designed until it is reliable. For this reason it is urgent that all engineers who design electronic equipment become familiar with all the problems of equipment reliability and the techniques for solving them, and to utilize this knowledge in all their work. Only then can mature, complete design of electronic equipment become a reality.

Reliability, What and Why is the first of a series of four articles written by Karle Packard. The next three articles will appear on these pages in May, June and July, 1959.

Part 2—The Nature of the Reliability Problem

Part 3—The General Methods of Solving the Reliability Problem

Part 4—The Language of Reliability

A complete bound set of our third series of articles is available on request. Write to Harold Hechtman at AIL for your set.

Industrial Engineering Notes*

Association Activities

The Formation of the National Stereophonic Radio Committee, a committee of technical experts to study methods of braodcasting stereophonic sound by AM, FM and TV stations, was announced by Dr. W. R. G. Baker, director of the EIA Engineering Department, as a service to the public, the FCC and the industry. Within the new committee, six panels will sift the technical factors involved in broadcasting stereo sound in compatible fashion; that is, in such a way as not to interfere with the existing public broadcast service. Existing receivers will continue to provide their present service; however, receivers specially built for stereo reception will be able to provide stereo sound from compatible stereo broadcast signals. Dr. Baker said all known technical proposals for such systems would receive careful study, and he urged all those who have such proposals to submit them to the Committee, through the associate director of the EIA Engineering Department, Virgil M. Graham, in New York City. The results of these studies will be made available to the public and to the FCC. Panels have been set up to study the application of proposed systems to transmitters and receivers, of existing and prospective designs, the problems of networks and other interconnecting facilities, field testing of signals to uncover factors influencing coverage and interference, and the subjective aspects of hearing involved in listening to stereo sound. A panel on system specifications will, when possible, formulate consistent sets of signal specifications for each form of broadcasting and provide an engineering evaluation of the performance implied in each set of specifications. The formation of such a committee was first suggested last November by Dr. Baker, who is vice-president of the Syracuse University Research Corp. and a retired vice-president of General Electric Co. Dr. Baker served as chairman of the National Television System Committees which formulated the technical standards underlying monochrome and color television. Chairman of the new NSRC Committee is C. Graydon Lloyd of the General Electric Co. George H. Brown of RCA is vice-chairman. A Coordinating Committee, consisting of the panel chairman, under Chairman Donald G. Fink, of Philco Corp., will oversee the panel assignments. The National Stereophonic Radio Committee will operate under the policy direction of an Administrative Committee, headed by Dr. Baker, with David B. Smith, of Philco Corp., as vice-chairman.

Following are the formal charter of the Committee, the panel designations and the scope of each. EIA emphasized that any technically qualified individual from any organization, whether or not associated with EIA, is welcome and invited to serve on the panels. Inquiries concerning membership and offers to serve should be addressed to Mr. Graham in New York. The charter of NSRC follows: "The function of the NSRC will be to make detailed technical studies of the several possible methods of providing compatible stereo sound for the AM, FM and TV broadcast services. The objective of these studies will be "1) To clarify the technical issues as between the several possible systems for each of these services. 2) To verify the technical conclusions through appropriate field tests and obtain such information as may be necessary for channel utilization purposes for the determination of the choice of standards. 3) To delineate appropriate signal specifications for the several services based upon the best scientific information and field test data available to the committee." The results of the studies will be made public as soon as the studies have been completed. The panel organization and scopes are as follows. *Coordination Committee:* The Coordination Committee shall consist of the chairman and vice-chairman of each Panel. This Committee shall coordinate the activities of the Panels, shall prepare definitions, and shall prepare NSRC Reports for the approval of the NSRC Operating Committee. *Panel 1—System Specifications:* Panel 1 shall consider system proposals for compatible stereophonic broadcasting; shall identify the technical issues in said proposals and refer them where necessary to other panels for detailed study; shall formulate a consistent set of transmission specifications for each form of broadcasting; and shall provide an over-all evaluation of the system performance implied in the specifications. *Panel 2—Interconnecting Facilities:* Panel 2 shall study and recommend technical characteristics of interconnecting lines, networks, studio-transmitter links and related stereotransmission facilities between program origination points and the transmitters proper, said characteristics to include tolerable limits on cross talk, relative time delay, frequency response, gain, and such other matters as must be controlled to assure a stereo signal of adequate quality at the transmitter input. *Panel 3—Broadcast Transmitters:* Panel 3 shall study the system proposals referred to it by Panel 1 with particular regard to 1) the feasibility of the proposed transmission method and 2) methods of adapting the proposals to existing broadcast transmitters. *Panel 4—Broadcast Receivers:*

(Continued on page 34A)

* The data on which these Notes are based were selected by permission from *Weekly Reports*, issues of January 26, February 2, 9 and 16, published by the Electronic Industries Association whose helpfulness is gratefully acknowledged.

Meetings with Exhibits

● As a service both to Members and the industry, we will endeavor to record in this column each month those meetings of IRE, its sections and professional groups which include exhibits.

April 16-18, 1959

SWIRECO, Southwestern IRE Regional Conference & Electronics Show, Dallas Memorial Auditorium & Baker Hotel, Dallas, Tex.
Exhibits: Mr. John McNeely, Southwestern Bell Telephone Co., 308 South Akard St., Dallas 1, Tex.

April 21-22, 1959

Spring Technical Conference on Electronic Data Processing, Engineering Societies Bldg., Cincinnati, Ohio
Exhibits: Mr. John I. Mika, Crosley Div., Avco Mfg. Co., 1329 Arlington St., Cincinnati 25, Ohio

May 4-6, 1959

National Aeronautical Electronics Conference, Dayton Biltmore Hotel, Dayton, Ohio.
Exhibits: Mr. Edward M. Lisowski, General Precision Lab., Inc., Suite 452, 333 West First St., Dayton 2, Ohio.

May 6-8, 1959

Seventh Regional Technical Conference and Trade Show, University of New Mexico, Albuquerque, N.M.
Exhibits: Mr. Earl C. Davis, P.O. Box 3262, Albuquerque, N.M.

June 3-5, 1959

Armed Forces Communications & Electronics Association Convention & Exhibit, Sheraton-Park Hotel, Washington, D.C.
Exhibits: Mr. William C. Copp, 72 West 45th St., New York 36, N.Y.

June 4-5, 1959

Third National Conference on Production Techniques, Villa Hotel, San Mateo, Calif.
Exhibits: Mr. Estrada Fanjul, Stanford Research Institute, Menlo Park, Calif.

June 13-22, 1959

International Conference on Information Processing, UNESCO House & Palais d'Exhibition, Paris, France.
Exhibits: Mr. E. M. Grabbe, Ramo Wooldridge Corp., Box 45067, Airport Station, Los Angeles 45, Calif.

June 29-July 1, 1959

Third National Convention on Military Electronics, Sheraton-Park Hotel, Washington, D.C.
Exhibits: Mr. L. David Whitelock, BuShips, Electronics Div., Dept. of Navy, Washington 25, D.C.

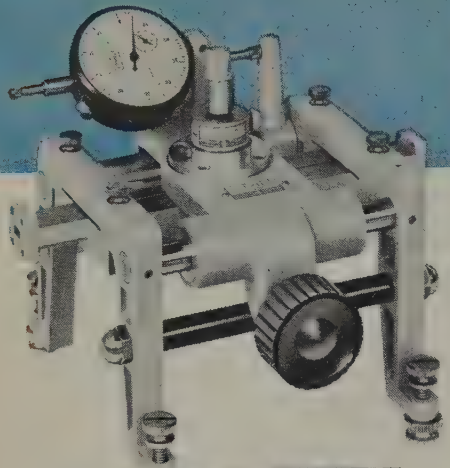
August 18-21, 1959

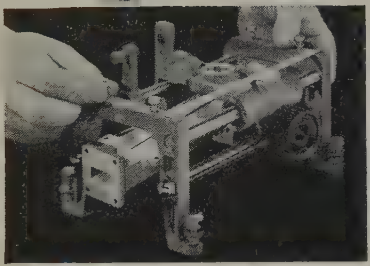
WESCON, Western Electronic Show and Convention, Cow Palace, San Francisco, Calif.
Exhibits: Mr. Don Larson, WESCON, 1435 La Cienega Blvd., Los Angeles, Calif.

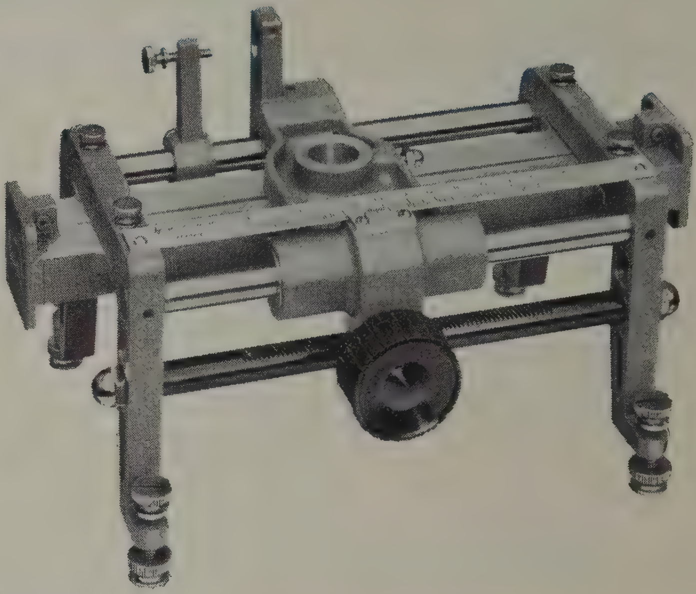
(Continued on page 12A)

Measure impedance and other system characteristics,

500 MC to 40 KMC







hp 809B and 814B
UNIVERSAL PROBE CARRIAGES

- **Coverage 3 to 40 KMC**
- **Sections interchange in 30 seconds**
- **Dial gauge accuracy, highest stability**

Models 809B and 814B are rugged, precision Universal Probe Carriages designed for use, respectively, with hp 810B and 815B waveguide slotted sections. The 809B/810B combination covers frequencies 3.95 to 18.0 KMC, and the 814B/815B combination covers frequencies 18.0 to 40.0 KMC. For waveguide measurements involving several bands, the cost of a special probe and carriage assembly for each band is eliminated and much engineering time is saved since waveguide sections can be changed in 30 seconds. Model 809B has a vernier scale reading to 0.1 mm and can be fitted with a dial gauge for greater accuracy. Model 814B is equipped with a dial indicator reading to 0.01 mm.

Specifications

hp 809B Universal Probe Carriage

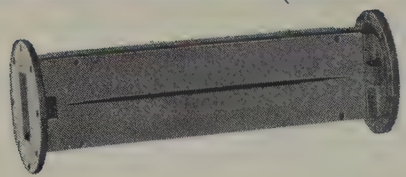
Carriage:	Mounts hp 810B Slotted Sections and hp 806B Coaxial Slotted Section (not shown: 3 to 12 KMC, 50 ohms impedance, Type N connectors).
Probe Required:	hp 442B Broadband Probe plus hp 440A Detector or hp 444A Untuned Probe.
Probe Travel:	10 centimeters.
Accuracy	With waveguide sections, 1.02 SWR easily read. Slope error eliminated by adjustment.
Price:	$160.00.

hp 814B Universal Probe Carriage

Carriage:	Mounts hp 815B Slotted Sections.
Probe Required:	hp 446B Untuned Probe.
Accuracy:	SWR of 1.02 easily read.
Price:	$200.00.

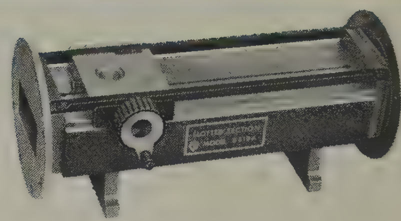
hp WORLD'S MOST COMPLETE LINE OF PRECISION,

quickly, accurately, with these low cost, precision instruments!



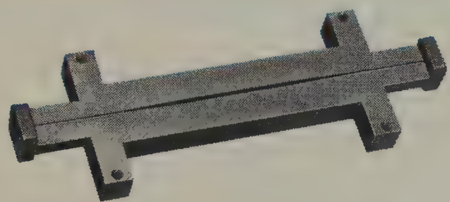
ⓗⓟ 810B Waveguide Slotted Sections— 3.95 to 18.0 KMC.

These accurately machined sections of waveguide have a small, tapered, longitudinal slot, and fit the 809B Universal Probe Carriage in a precisely indexed position. A traveling probe mounted on the carriage samples the electric field along the slot, and permits precise plotting of variations. Slot reflection is less than 1.01 SWR. For prices, list of 810B waveguides available, see Table 1 below.



ⓗⓟ S810A Waveguide Slotted Section— 2.6 to 3.95 KMC.

This instrument is a conventional slotted waveguide complete with a probe carriage mounted directly on the section. It is available in the S-band only and will operate with ⓗⓟ 442B or 444A probes. SWR less than 1.01. $450.00.

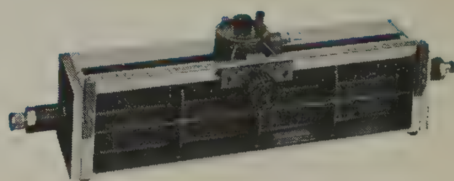


ⓗⓟ 815B Waveguide Slotted Sections— 18.0 to 40.0 KMC.

Available in K and R band sizes, these waveguide slotted sections are similar to 810B sections and, like 810B units, are accurately machined from precision castings to insure a uniform cross-section. Prices and details below.

Table 1—ⓗⓟ 810B/815B Slotted Sections.

Model	Frequency Range KMC	Fits Waveguide Size (in.)	Overall Length (in.)	Price
G810B	3.95 - 5.85	2 x 1	10¼	$110.00
J810B	5.20 - 8.20	1½ x ¾	10¼	110.00
H810B	7.05 - 10.0	1¼ x ⅝	10¼	110.00
X810B	8.20 - 12.4	1 x ½	10¼	90.00
P810B	12.4 - 18.0	.702 x .391	10¼	110.00
K815B	18.0 - 26.5	.500 x .250	4½	265.00
R815B	26.5 - 40.0	.360 x .220	4½	265.00



ⓗⓟ 805A/805B Coaxial Slotted Lines— 500 MC to 4 KMC.

For SWR, wavelength, impedance and system flatness measurements in coaxial systems. Exclusive ⓗⓟ parallel-plane design for higher accuracy, stability. Negligible slope, SWR less than 1.04, reads in cm and mm to 0.1 mm. ⓗⓟ 805A, for 50 ohm lines, Type N connectors, $450.00. ⓗⓟ 805B, for 46.3 ohm lines, UG-45/U male and UG-46/U female connectors, $450.00.

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Meetings with Exhibits

(*Continued from page 8A*)

September 23-25, 1959
Special Technical Conference on Non-Linear Magnetics and Magnetic Amplifiers, Shoreham Hotel, Washington, D.C.
Exhibits: Mr. S. Lax, G-L Electronics Co., Inc., 2921 Admiral Wilson Blvd., Camden 5, N.J.

September 28-30, 1959
National Symposium on Telemetering, Civic Auditorium & Whitcomb Hotel, San Francisco, Calif.
Exhibits: Mr. Robert A. Grimm, Dymec, Inc., 395 Page Mill Road, Palo Alto, Calif.

October 5-7, 1959
Fifth National Communication Symposium, Hotel Utica, Utica, N.Y.
Exhibits: Mr. E. William Morris, 224 Fairway Drive, New Hartford, N.Y.

October 7-9, 1959
IRE Canadian Convention, Exhibition Park, Toronto, Ont., Canada.
Exhibits: Mr. F. G. Heath, IRE Canadian Convention, 1819 Yonge St., Toronto 7, Ont., Canada.

October 12-14, 1959
National Electronics Conference, Hotel Sherman, Chicago, Ill.
Exhibits: Mr. Arthur H. Streich, National Electronics Conference, Inc., 228 N. La Salle St., Chicago 1, Ill.

October 26-28, 1959
East Coast Aeronautical & Navigational Electronics Conference, Lord Baltimore Hotel & 7th Regiment Armory, Baltimore, Md.
Exhibits: Mr. R. L. Pigeon, Westinghouse Electric Corp., Air Arm Div., P.O. Box 746, Baltimore, Md.

November 9-11, 1959
Fourth Instrumentation Conference, Atlanta, Ga.
Exhibits: Dr. B. J. Dasher, School of E.E., Georgia Institute of Technology, Atlanta 13, Ga.

November 30-December 3, 1959
Eastern Joint Computer Conference, Hotel Statler, Boston, Mass.
Exhibits: Mr. John M. Broomall, Burroughs Corporation, Paoli, Pa.

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Note on Professional Group Meetings: Some of the Professional Groups conduct meetings at which there are exhibits. Working committeemen on these groups are asked to send advance data to this column for publicity information. You may address these notices to the Advertising Department and of course listings are free to IRE Professional Groups.

IRE News and Radio Notes

Calendar of Coming Events and Authors' Deadlines*

1959

Nuclear Cong., Cleveland, Ohio, Apr. 5–10

4th Conf. on Industrial Instrumentation & Control., Ill. Inst. Tech., Chicago, Ill., Apr. 14–15

SWIRECO (Southwestern Regional Conference), Dallas, Texas, Apr. 16–17.

Conf. on Analog and Digital Recording and Controlling Instrumentation, Bellevue-Stratford Hotel, Phila., Pa., April 20–21

Spring Tech. Conf. of Cincinnati Sec. of the IRE, April 21, 22.

115th Ann. Electrochemical Soc. Meeting, Hotel Sheraton, Phila., Pa., May 3–7

Nat'l Aero Elec., Conf. Biltmore and Miami-Pick Hotel, Dayton, Ohio, May 4–6

Fifth Annual Flight Test Instr. Symp., Seattle, Wash., May 4–7

Interdisciplinary Conf. on Self-Organizing Systems, Mus. Sci. & Indus., Chicago, Ill. May 5–6

URSI Spring Meeting, Washington, D. C., May 5–7

Elec. Components Conf., Ben Franklin Hotel, Philadelphia, Pa., May 6–8

7th Reg. Tech. Conf. and Trade Show, Univ. of N. M., Albuquerque, N. M., May 6–8

Symp. on Radar Return, Univ. of N. Mex., May 11–12

Joint Conf. on Auto. Tech., Pick-Congress Hotel, Chicago, Ill., May 11–13

Internat'l Conv. on Transistors and Associated Semiconductor Devices, Earls Court, London, May 25–29

Australian IRE Radio Eng. Conv., Univ. of Melbourne, Victoria, Aus., May 25–30

Internat'l Conf. on Med. Elec., Paris, France, June

Microwave Theory & Tech., 1959 Nat'l Symp., Harvard Univ., Cambridge, Mass., June 1–3 (DL*: Feb. 15, Dr. H. J. Riblet, 92 Broad St., Wellesley, Mass.)

Prod. Tech. Symp., Villa Hotel, San Mateo, Calif., June 4–5

Symp. on Electromagnetic Theory, Univ. of Toronto, Toronto, Can., June 15–20

Int'l Conf. on Info. Processing, UNESCO House, Paris, France, June 15–20

Int'l Symp. on Circuit & Information Theory, Univ. of Calif. at Los Angeles, Los Angeles, Calif., June 16–18.

6th Annual Meeting of Soc. of Nuclear Med., Palmer House, Chicago, Ill. June 18–19.

AIEE Conf. on Air Transportation, Olympic Hotel, Seattle, Wash., June 24–26

* DL = Deadline for submitting abstracts.

(Continued on page 15A)

PGAC Schedules Conference

The IRE Professional Group on Automatic Control will sponsor a National Automatic Control Conference on November 4–6, 1959, in Dallas, Texas at the new Sheraton-Dallas Hotel. Control groups from other organizations such as the PGIE, AIEE, ASME, and ISA will participate in the activitis.

Although the deadline for papers will not occur until June 1, 1959, four copies of summaries should be submitted as soon as possible to G. S. Axelby, Westinghouse Electric Corp., Box 746, Baltimore 3, Md.

In order to facilitate selection of papers, the 1000 to 1500 word summary must 1) stage clearly what has been accomplished; 2) indicate whether (a) the material is primarily theoretical or experimental, (b) practical applications are included, and (c) the paper is believed to be an original contribution or an extension of an earlier paper; and 3) include a pertinent bibliography.

Accepted papers will be published in the PGAC Transactions.

Brewer Receives Reliability and Quality Control Award

At the Fifth National Symposium on Reliability and Quality Control, held in Philadelphia in January, Mr. Ralph Brewer of the Research Labs. of the General Electric Co. Ltd. of England was given the 1958 National Reliability Award for his paper entitled "Life Tests of Electron Tubes and the Analysis of Failure Causes," which was read at the National Symposium in 1958. The award is made to the author of the best technical paper presented at this annual event which is jointly sponsored by the IRE, the American Institute of Electrical Engineers, the American Society for Quality Control, and the Electronic Industries Association. Mr. Brewer's paper was the only overseas contribution in the total of fifty papers read.

MAECON—1959 Calls for Papers

The eleventh Annual Mid-America Electronics Conference (MAECON—1959) will be held this year in the spacious Kansas City Municipal Auditorium and Hotel Muehlebach on November 3, 4, and 5.

Technical papers are wanted on engineering management, engineering education, astronautics, electro-magnetics, nucleonics, airborne electronics, medical electronics, and other electronic subjects. Abstracts of approximately 100 words should be sent not later than July 1 to Dr. Sheldon L. Levy, Midwest Research Institute, 425 Volker Blvd., Kansas City 10, Mo.; or Dr. Charles Halijak, Kansas State College, Manhattan, Kansas.

Inquiries concerning exhibition booth space should be sent to Mr. John V. Parks, Exhibits Chairman, Bendix Aviation Corp., P.O. Box 1159, Kansas City 41, Mo.

PGVC Call for Papers

The 10th National Conference of the IRE Professional Group on Vehicular Communications is seeking material for their technical program to be held in St. Petersburg, Fla., on December 3–4, 1959. Papers on subjects covering vehicular systems and equipment designs and discussions on new or unusual system techniques, applications of new types of components or related circuitry, interference reduction or bandwidth utilization are invited. Emphasis on field proven subjects is desired. Manufacturers and users are urged to take advantage of this opportunity to share their experiences in this rapidly expanding communications field.

An abstract of 500 words is required for review by the papers committee by June 30, 1959. These should be mailed to J. R. Neubauer, RCA, Bldg 1–4, Camden 2, N. J. Authors will be notified by August 15 of their acceptance.

Top award winners and officials confer at the Fifth National Symposium on Reliability and Quality Control at the Bellevue Stratford Hotel in Philadelphia, January 13–14, 1959. Shown in the picture (*from left to right*) are William MacCrehan, Bendix Radio; Dr. W. H. Pickering, Director of Jet Propulsion Labs., California Inst. of Tech.; C. M. Ryerson, Symposium General Vice-Chairman; David R. Hull, EIA President; P. K. McElroy, PGRQC Chairman; and Ralph Brewer, General Electric, LTD.

Call for Papers by the 1959 National Symposium on Telemetering

The 1959 National Symposium on Telemetering, sponsored by the IRE Professional Group on Space Electronics and Telemetry and to be held in San Francisco, Calif., September 28–30, 1959, has issued a call for papers. Prospective authors are invited to submit papers on significant work in telemetry and related fields, such as the following: 1) space vehicle communication systems; 2) satellite instrumentation; 3) data processing equipment; 4) new developments in telemetry equipment; 5) flight test electronics.

In addition to papers on new developments, tutorial papers having special interest to the professional group are desired. As in 1957 and 1958 awards will be given for outstanding papers.

It is planned to distribute the Symposium Transactions in advance of the meeting. In order to allow sufficient time for author selection, preparation, and publication of papers, strict adherence to the following schedule will be necessary.

May 10—Deadline for receipt, in triplicate, of a 100–200 word abstract and a 500-word summary of the paper.
June 1—Notification of paper selection or rejection.
July 10—Deadline for receipt of papers.

All correspondence should be addressed to George L. Larse, Program Chairman, Lockheed Aircraft Corp., Missile Systems Div., Sunnyvale, Calif.

Exhibitors: San Francisco's Civic Auditorium provides excellent NST facilities. For space information write R. A. Grimm, Dymec, Inc., 395 Page Mill Rd., Palo Alto, Calif.

WESCON Announces Plans

Authors wishing to present papers at the 1959 Western Electronic Show and Convention technical sessions, to be held in San Francisco, August 18–21, must submit them by May 1. Required are 100–200 word abstracts, together with complete texts or additional detailed summaries, which should be sent to the Chairman of the Technical Program, Dr. Karl R. Spangenberg, c/o WESCON, 60 West 41st Ave., San Mateo, Calif.

This year WESCON is planning three important innovations to upgrade and enliven the technical session presentations and discussions: 1) The technical program will comprise the usual 40 daytime sessions, but with only three full-length papers in each. 2) A panel of two or three experts will be invited to comment at the conclusion of each paper. 3) The IRE WESCON CONVENTION RECORD will be available at the convention. Convention authors will be expected to submit complete manuscripts by July 1, prepared for the Record in accordance with special instructions which will be sent at the time the paper is accepted.

Authors will be notified of acceptance or rejection by June 1.

Gen. Kepner Heads Military Electronics Meeting

Lt. Gen. William Kepner, USAF (Ret.), Chairman of the Board of Radiation, Inc., Orlando, Fla., is Convention President for the Third National Convention on Military Electronics—1959, to be held in Washington, D. C., June 29–July 1. It is sponsored by the Professional Group on Military Electronics of the IRE. At the 1958 Convention, also held in Washington, more than 3300 engineers, scientists, and executives from industry, Government agencies and laboratories, the Armed Forces, universities and embassies listened to more than 100 technical papers and looked at 71 exhibits of new developments in the field of military electronics. Included in this branch of electronics are such major topics as satellite electronics; space navigation; guidance and control systems; electronic propulsion; reconnaissance systems; simulation; and communication systems.

General Kepner, who has had a 32-year career with the Air Force, holds the aeronautical ratings of command pilot, combat observer, senior balloon pilot, Zeppelin pilot, semi-rigid pilot, and metalclad airship pilot. He is a veteran of both World Wars, and during his command of various tactical and strategic forces in Europe in World War II, participated in 24 missions over hostile areas.

Gen. Kepner has been awarded the Distinguished Service Cross, Purple Heart, Distinguished Flying Cross, Legion of Merit with two Clusters, Distinguished Service Medal with Cluster, Bronze Star Medal, Air Medal, and USMC Good Conduct Medal. He has also been awarded numerous foreign decorations including the British Order of Commander of the Bath, Belgian Order of Crown with gold palm; Commander of French Legion of Honor, and French Croix de Guerre with palm.

In 1950, he was named Commander-in-Chief of the Alaskan Command at Fort Richardson, Alaska. He also has been chief of the Atomic Energy Division in the office of the Deputy Chief of Staff for Research and Development, as well as Assistant Deputy Chief of Staff, Operations, for Atomic Energy.

Lt. Gen. William Kepner, USAF (Ret.), Convention President for the Third National Convention on Military Electronics.

Calendar of Coming Events and Authors' Deadlines*

(*Continued from page 14A*)

Nat'l Conv. on Mil. Elec., Sheraton Park Hotel, Washington, D. C., June 29–July 1.

Denver Res. Inst. Sixth Annual Symp. on Computers and Data Processing, Stanley Hotel, Estes Park, Colo. July 30, 31

Natl. Ultrasonics Symp., Stanford Univ., Stanford, Calif., Aug. 17

WESCON, San Francisco, Calif., Aug. 18–21

PGEWS Symp., Boston and Los Angeles, Sept. 17–18 (A.M. Cross, Raytheon Mfg. Co., Wayland, Mass.; J. M. Cryden, Hughes Aircraft Corp., Culver City, Calif.)

Special Tech. Conf. on Nonlinear Magnetics and Mag. Amplf., Shoreham Hotel, Washington, D.C., Sept. 23–25

Nat'l Symp. on Telemetering, Civic Aud. & Whitcomb Hotel, San Francisco, Calif., Sept. 28–30 (DL* May 10, G. Larse, Missile Syst. Div., Lockheed Aircraft Corp., Sunnyvale, Calif.)

Symp. on Indus., Elec., Mellon Inst., Pittsburgh, Pa., Sept. 30–Oct. 1

5th Aeronautical Comm. Symp., Hotel Utica, Utica, N.Y., Oct. 5–7

5th Nat. Communications Symp., Hotel Utica, Utica, N.Y. Oct. 5–7.

IRE Canadian Conv., Toronto, Can., Oct. 7–9

Nat'l Elec. Conf., Sherman Hotel, Chicago, Ill., Oct. 12–15

(DL*: May 1, M.E. Van Valkenburg, Dept. of E.E., Univ. of Illinois, Urbana, Ill.

East Coast Conf. on Aero. and Nav. Elec., Baltimore, Md., Oct. 26–28

Electron Devices Mtg., Shoreham Hotel, Washington, D. C., Oct. 29–31

Mid. Amer. Elec. Conv., Kansas City, Mo., No. 3–4

Nat'l Conf. on Automatic Control, New Sheraton Hotel, Dallas, Tex., Nov. 4–6

Radio Fall Mtg., Syracuse, N. Y., Nov. 9–11

4th Conf. on Instrumentation, Atlanta, Ga., Nov. 9–11

12th Ann. Conf. on Elec. Tech. in Med. & Bio., Penn Sheraton Hotel, Phila., Pa., Nov. 10–12

1959 NEREM (Northeast Electronics Res. & Engng. Meeting), Boston Commonwealth Armory, Boston, Mass., Nov. 17–19

Eastern Joint Comp. Conf., Hotel Statler, Boston, Mass., Nov. 30–Dec. 3

4th Midwest Symp. on Circuit Theory, Brooks Mem. Union, Marquette Univ., Milwaukee, Wisc., Dec. 1–2. (DL* May 1, J. J. Graham, 1515 W. Wisc. Ave., Milwaukee 3, Wis.)

PGVC Annual Meeting, St. Petersburg, Fla., Dec. 3–4 (DL*: Jun 30, J. R. Nubauer RCA Bldg. 1–4 Camden, N.J.)

4th Midwest Symp. on Circuit Theory, Marquette Univ., Milwaukee Wisc., Dec. 1–2 (DL*: May 1, S. Krupnik, Jr., E.E. Dept., Marquette Univ., Milwaukee, Wisc.)

* DL = Deadline for submitting abstracts.

Members of the 1959 IRE Canadian Convention Executive Committee: (*seated, left to right*) A. P. H. Barclay, IRE Region 8 Director; H. W. Jackson, Recording Secretary; Eric L. Palin, General Chairman; Fred J. Heath, Vice-Chairman; Grant Smedmor, Convention Manager; (*standing left to right*) D. K. Ritchie; Chairman Technical Program; S. F. Love, Exhibit Award Committee; G. Armitage, Chairman Exhibits Committee; R. G. Bullock, Chairman Social Activities Committee, H. F. Shoemaker, Chairman IRE Toronto Section; T. Purdy, Region 8 Liaison; A. Jones, Chairman Advertising and Publicity Committee; R. M. Lynd, Chairman Finance Committee; L. M. Price, Chairman Registration and Reception Committee.

IRE Canadian Convention Scheduled for October

The fourth annual IRE Canadian Convention and Exposition will be held in the Automotive Building, Exhibition Park, Toronto, on October 7–9, it was announced by Eric L. Palin, General Chairman of the Executive Committee.

On the first day of the 1959 Convention a Fall Executive meeting of the IRE International will be held in Toronto. It will be followed on the second and third days by a Board meeting of the organization. Twenty directors and officers, including the eight Regional Directors, are expected to participate in these executive sessions. A. P. H. Barclay, IRE Region 8 Director is making arrangements for the Toronto meetings which will result from the IRE International's acceptance of a long standing invitation from the Canadian Convention Committees.

Don K. Ritchie, Chairman of the Technical Program, is now inviting IRE members and other scientists to submit papers for possible use during the 1959 Convention. These may be on any subject in the fields of nucleonics and electronics. Interested authors should first submit, in duplicate, 500 word summaries of their papers to Mr. D. K. Ritchie, IRE Canadian Convention Offices, 1819 Yonge Street, Toronto, Canada. One hundred word abstracts of papers are also required for use in printed booklets covering the technical program.

The three previous IRE Canadian Conventions have been Canada's largest annual scientific events. Last year's attracted over 10,000 scientists, engineer's technicians, and businessmen. During the three-day Technical Program 117 scientific papers in 23 categories ranging from medical and industrial electronics to cosmic rays and communications were presented by Canadian, U. S. and overseas authors. In the 1958 Exposition over 150 organizations, including several U. S. and overseas compan'es, used 300 exhibit booths to show electronic and nucleonic equipment. To broaden the scope of the Exposition in 1959 the Canadian Convention is sending invitations to scientific agencies of overseas governments as well as overseas commercial firms in the nucleonic and electronic fields. Another highlight of the 1959 Convention will be the engagement of a headline guest speaker for the Convention Banquet in the Royal York Hotel.

Members of the 1959 IRE Canadian Convention Executive Committee are as follows: E. L. Palin, General Chairman; R. G. Bullock, Chairman, Social Activities Committee; H. W. Jackson, Recording Secretary; R. M. Lynd, Chairman, Finance Committee; H. F. Shoemaker, Chairman, IRE Toronto Section; E. Jones, Chairman, Advertising and Publicity Committee; F. J. Heath, Vice-Chairman; D. K. Ritchie, Chairman, Technical Program; S. F. Love, Exhibit Award Committee; G. Armitage, Chairman, Exhibits Committee; L. M. Price, Chairman, Registration and Reception Committee; H. R. Smyth, Chairman, Exhibit Award Committee; G. Smedmor, Convention Manager; A. P. H. Barclay, IRE Region 8 Director; T. Purdy, Region 8 Liaison; Dr. G. Sinclair, past General Chairman.

Solid-State Conference Publishes Papers Digest

A digest of the technical papers presented in February at the 1959 Solid-State Circuits Conference in Philadelphia, has been published and is available at $4.00 per copy from Henry G. Sparks, Moore School of Electrical Engineering, Univ. of Pennsylvania, Philadelphia, Pa. The Digest contains informative 650-word summaries and illustrations of all conference papers, and includes an author index, program schedule, and space for making notes. For a listing of papers, see the program on page 22A of the January, 1959 issue of Proceedings.

Summer Program Announced by M.I.T.

A two-week summer program at the Massachusetts Institute of Technology on "Finite- and Infinite-State Machines," will be given August 3–14, 1959.

This two-week summer program will constitute an introduction to the logical aspects of finite- and infinite-state machines. The material to be covered will range from the design and analysis of sequential machines to the theory of computability and Turing machines.

Design procedures which eliminate undesirable effects in asynchronous finite-state machines will be illustrated. The limitations and capabilities of both finite- and infinite-state machines will be characterized. Other topics will include the generation of pseudorandom sequence by linear finite-state machines and canonical forms for information-lossless circuits. The present status of and directions for future research in multidimensional logical arrays, use of unreliable components in the design of logical machines, and computation in the presence of noise will be explored.

The program will be directed by Professors Dean Arden and David Huffman of the Department of Electrical Engineering. Other lecturers will include Professors Edward Arthurs, S. H. Caldwell, Peter Elias, Dr. Belmont Farley, Mr. Fred Hennie, and Professors Marvin Minsky and Hartley Rogers.

Application forms and full information may be obtained from Dr. James M. Austin, Director of the Summer Session, Room 7-103, M.I.T., Cambridge 39, Mass.

Army MARS April Schedule

The First U. S. Army MARS SSB Technical Network, operating each Wednesday at 9 P.M. New York time (whether EST or EDT) on 4030 kilocycles upper sideband, announces the speaker schedule for April.

Apr. 1—"Variable Reactance (Parametric) Amplifiers," Dr. S. Deutsch, Assoc. Prof. of Elec. Engng., Polytechnic Inst. of Brooklyn.

Apr. 8—"Electro-Mechanical Filters," O. P. Olson, Dept. Head, Res. and Dev., Collins Radio Co.

Apr. 15—"Phosphors and Electro-Luminescense," Dr. P. Goldberg, Engineering Specialist, Sylvania Elec. Prods. Corp.

Apr. 22—"Atlas-Score Communications System," by S. P. Brown, Deputy Director, Transmission Facilities Div., U. S. Army Res. and Dev. Lab., Fort Monmouth, N. J.

Apr. 29—"Interchanging Scientific Information by Multilateral Radio Communication," S. E. Piller, W2KPQ/A2KPQ, Director, First U. S. Army MARS SSB Tech.

The First Army Technical Network will pass another milestone when on April 8 the Collins Radio Co. speaker, originating from California, will be broadcast throughout the eastern United States. This will be accomplished by means of a transcontinental phone patch between Burbank and New York City.

Sir Francis Drake Cracks a Case

One foggy day in 1588, a single ship of the Spanish Armada managed to sneak behind Drake's entire British fleet lying in the English Channel off Plymouth Hoe, and drop a 10 pounder smack in the middle of a bowling match between Sir Francis and his friend Walter Raleigh.

The new radar was caught completely by surprise. Had the IFF (Identification: Francis or Foe?) system failed? Was the operator tuned to the wrong Channel? Was there something wrong with the tubes? Drake was determined to find out. He was inside the shack in a trice, whatever that is. "Avast!" he roared at the radarman, "I must inspect those tubes!"

Drake picked up a magnetron and looked at it. "Aha!" he exclaimed. Just as my razor-keen mind suspected! With that, he seized the hapless operator by the throat and shook him like a tumblerful of sidecars. "I arrest you for stealing Bomac tubes and substituting these inferior substitutes, WILLIAM SHAKESPEARE!"*

"I confess, how'd you guess?" said Shakespeare, ever the poet.

"Elementary for a razor-keen mind like mine," answered Drake. "Only you could have conceived the cunning scheme of replacing Bomac tubes with factory seconds labeled "Bethmac" as a publicity stunt for your new play — Macbeth!"

"Yours is a razor-keen mind indeed!" marveled Shakespeare as they led him away. "I haven't even written Macbeth yet!"

"Plenty of time where you're going," said Drake — and went off to bowl over the Armada.

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Symposium on Radar Return

A symposium will be held at the University of New Mexico, sponsored by the Naval Ordnance Test Station, China Lake, Calif., on May 11–12. Papers dealing with radar return from ground and water targets will be presented. Subjects having application to radar altimeters are to receive special emphasis.

The sessions on May 11 will include papers that are unclassified and are open to all interested in the field of radar return. Classified sessions requiring a clearance of Secret will be held on May 12. The meetings will be held in Johnson Gymnasium at the University. Please make clearance arrangements with Mr. H. L. Walker, Director of Research, Box 4036, Albuquerque, New Mex.

Weber Speaks to Washington Section

In February Dr. Ernst Weber, IRE President and President of the Polytechnic Institute of Brooklyn, spoke at the annual banquet of the IRE Washington Section about the present need for "scientific engineers," rather than engineers equipped only with skills. He emphasized the importance of attracting outstanding engineering students into graduate schools and urged that the education of scientists and engineers include a "significant and meaningful exposure" to philosophy, literature, art, economics and government. At the same time he added that the liberal arts curriculum should include the study of scientific methods, perhaps in terms of history of science followed by real study in a particular branch of science. In addition, he declared, "Unless we can secure adequate outside support for our basic research programs either from government or industry, or both, we shall not be able to maintain the high level of graduate teaching and of research which is so indispensable to faculties in graduate schools."

Other events of the Banquet included the presentation of the 1959 Student Awards to ten outstanding students in the Washington area. IRE Fellow awards were also presented to Washington Section members M. Apstein, R. Bateman, J. N. Bridges, T. M. Odarenko, R. D. Parker, and J. D. Wallace.

Dr. R. M. Page, Director of Research of the U. S. Naval Research Lab. and Chairman of the section, presented "Patron Awards" for distinguished service to the section to G. P. Adair, Consulting Engineer; E. K. Jett, Vice-President, WMAR-TV; L. C. Smeby, Consulting Engineer; and H. W. Wells, Carnegie Institution of Wash.

Student award winners were R. M. Kieth, P. Cudmore, both of George Washington Univ.; S. P. Tynes, J. W. Breedlove, both of Howard Univ.; J. V. Popolo, P. W. Fish, both of Catholic Univ.; C. L. Gillis, R. E. Davis, both of the Univ. of Maryland; and R. F. Tuttle and E. P. Sheetz, both of Capitol Radio Engineering Institute.

The speakers at the Annual Banquet of the Washington Section are (*left to right*) R. I. Cole, Director, Third Region; Dr. R. M. Page, Chairman, Washington Section; Dr. E. Weber, President, IRE; and Rear Adm. R. Bennett, USN, Chief of Naval Research, toastmaster.

Air Force MARS Lists Broadcast Schedule

The Air Force MARS Eastern Technical Network, which broadcasts every Sunday from 2–4 P.M. (EST) on 3295 and 7540 kc, announces the following schedule.

April 5—"Comparison of Analog and Digital Computers," H. W. Merrihew, Supervisor of Course Preparation, Philco Technological Center.

April 12—"Characteristics of Transistorized Digital Computers," J. A. Maddox, Engng. Sec. Manager, Computer Lab., Philco, Corp.

April 19—"Installation and Maintenance of Radio Teletype," R. M. Clemick, Chief Instructor, Radio Communications Sec., Philco Technological Center.

April 26—"Physiological and Psychological Effects of Air Ionization," F. P. Speicher, Chief Biologist, Advanced Study Group, Philco Corp.

Kelly Award to be Given by AIEE

The Bell Telephone Laboratories and the American Institute of Electrical Engineers have announced the establishment of an award for achievement in the field of telecommunications, to be known as the Mervin J. Kelly Award.

The award is named in honor of Dr. Kelly, formerly president and now chairman of the board of Bell Laboratories, who retired on March 1 after 41 years of scientific and administrative service with the Bell Telephone System. Dr. Kelly is a Fellow of the AIEE and has been active in Institute affairs for 33 years. He is internationally recognized as one of the world's leading scientists and research administrators, whose distinguished achievements have played a large part in the advance of telecommunications during the past four decades.

The Kelly Award will be made annually by the AIEE to an individual who has made an outstanding contribution to the advancement of the art of telecommunications. It will consist of a bronze medal, a cash prize of $1000, and a certificate. The first award will be made in 1960. The award is being sponsored by Bell Laboratories but will be administered solely by the Institute.

Professional Group News

On February 10 the IRE Executive Committee approved the Huntsville Chapter of the Professional Group on **Space Electronics and Telemetry.**

Obituary

Lynde P. Wheeler (F'28), a past President of the IRE, died recently at the age of 84.

Dr. Wheeler was born in Bridgeport, Connecticut, on July 27, 1874. He received the Ph.B. degree in 1894 and the Ph.D. degree in 1902 from Yale University, where he then remained to teach until 1926.

From 1926 to 1936 he served as consultant at the U. S. Naval Research Laboratory. In 1936 he joined the Federal Communication Commission to organize the Technical Information Division, which he directed until 1946. From 1946 Dr. Wheeler served first as vice-president and later as staff consultant at Packard and Burns, Inc. in Needham, Massachusetts until his retirement in 1957.

An active member of the IRE, Dr. Wheeler served as President in 1943 and Director from 1940–1945. He was Chairman of the Institute Executive Committee in 1943 and at various times a member of the Bibliography, Board of Editors, Frequency Modulation, Meetings and Papers, Papers Procurement, Papers, Regular Papers, Membership Solicitation Policy, Nominations, Public Relations, and Standardization Committees. He was Chairman of the Washington Section of the IRE in 1930–1932.

He was a Fellow of the IRE, the American Physical Society, and the American Association for the Advancement of Science.

1959 IRE NATIONAL CONVENTION RECORD

The following important changes have been made this year with regard to the IRE NATIONAL CONVENTION RECORD:

1) Prices have been reduced by more than 50 per cent.
2) A special reduced rate has been established for members of the IRE Professional Groups.
3) The practice of distributing free copies to Professional Group members has been discontinued.

As in the past, the IRE NATIONAL CONVENTION RECORD, containing all available Convention papers, will be published in ten parts in July, 1959.

Professional Group members are entitled to purchase the Part sponsored by the Professional Group to which they belong at the special PG rate indicated below. Other Parts may be purchased at the IRE Member rate. There will be no free distribution.

IRE members may purchase any Part at the IRE member rate indicated below. However, if a member applies for membership in the appropriate Professional Group at the time he places his order, he will be entitled to the PG rate.

Nonmembers and libraries may place orders at the Nonmember and Library rates, respectively. Individuals who apply for IRE membership at the time they place their orders are entitled to the IRE Member rate.

Subscription agencies are entitled to the library rate.

Send orders, with remittance, to the Institute of Radio Engineers, Inc., 1 East 79 Street, New York 21, N. Y. Orders must be received by April 30, 1959.

IRE NATIONAL CONVENTION RECORD

Part	Sessions	Subject and Sponsoring IRE Professional Group	Prices for Members of Sponsoring Professional Group (PG), IRE Members (M), Libraries (L), and Nonmembers (NM)			
			PG	M	L	NM
1	38, 46, 53	Antennas & Propagation	$0.70	$1.05	$2.80	$3.50
2	34, 41, 49	Circuit Theory	0.70	1.05	2.80	3.50
3	8, 16, 23, 32, 39	Electron Devices Microwave Theory & Techniques	1.00	1.50	4.00	5.00
4	1, 9, 17, 25, 33, 40, 48	Automatic Control Electronic Computers Information Theory	1.20	1.80	4.80	6.00
5	7, 15, 24, 28, 36, 43, 51	Aeronautical & Navigational Electronics Military Electronics Space Electronics & Telemetry	1.20	1.80	4.80	6.00
6	6, 22, 27, 31, 35, 42, 44, 50	Component Parts Industrial Electronics Production Techniques Reliability & Quality Control Ultrasonics Engineering	1.40	2.10	5.60	7.00
7	11, 12, 19, 20, 26, 52	Audio Broadcast & TV Receivers Broadcasting	1.00	1.50	4.00	5.00
3	2, 4, 30, 37	Communications Systems Radio Frequency Interference Vehicular Communication	0.80	1.20	3.20	4.00
9	10, 14, 18, 21, 45, 47, 54	Human Factors in Electronics Instrumentation Medical Electronics Nuclear Science	1.20	1.80	4.80	6.00
10	3, 5, 13, 29	Education Engineering Management Engineering Writing & Speech	0.80	1.20	3.20	4.00
		Complete Set (10 Parts)	$10.00	$15.00	$40.00	$50.00

Industrial Instrumentation and Control Conference

April 14–15, 1959, Smith-Olson Auditorium, Chicago, Ill.

Tuesday Morning, April 14

Chairman, *Dr. R. B. Jacobs, Man., Engng. Res. Dept. Standard Oil Co., Whiting, Ind.*

Welcome Address, *Dr. C. E. Barthel, Jr., Asst. Dir., Armour Res. Found.*

"Evolution of Automatic Process Control," *G. A. Pettit, Barber-Coleman Co., Wheelco Div., Rockford, Ill.*

"New Developments in Stream Analysis," *V. H. Adams, Consol. Electrodynamics Corp., Pasadena, Calif.*

"Nuclear Magnetic Resonance Applications," *Dr. H. Rubin, Sen. Eng., Schlumberger Well Surveying Corp., Ridgefield, Conn.*

Tuesday Afternoon

Chairman, *Dr. S. Hori, Sen. Eng., Armour Res. Found.*

"Recent Developments in Transducer Technology," *Dr. Y. T. Li, Assoc. Prof., Aeronautic Eng., M.I.T., and President, Dynamic Instruments Co., Cambridge, Mass.*

"Use of Infrared Techniques in Industrial Instrumentation," *G. F. Warnke, President and H. L. Berman, Vice-President, Radiation Electronics Corp., Skokie, Ill.*

"Non-Destructive Eddy-Current Testing," *Dr. G. O. McClurg, Dir. Res., Magnaflux Corp., Chicago, Ill.*

Wednesday Morning, April 15

Chairman, *Dr. E. A. Keller, Staff Scientist, Motorola, Chicago Mil. Elec. Ctr., Chicago, Ill.*

"Electronic Photography," *M. L. Sugarman, Jr., Dir. Electrophotographic Res., American Photocopy Equipment Co., Chicago, Ill.*

"Applying Military Reliability Research to Industrial Electronics," *H. L. Wuerffel, Man., Rel. Anal. and Meas. Eng., Defense Electronics Prods., RCA, Camden, N. J.*

"Future Trends in Instrumentation," *D. J. McCowell, Branch Ind. Man., Minneapolis-Honeywell Regulator Co., Milwaukee, Wis.*

Wednesday Afternoon

Chairman, *R. E. Zenner, Vice-President, Union Thermoelectric Corp., Evanston, Ill.*

"Digital Control Systems—Present and Future," *Dr. M. Phister, Jr., Dir. Eng., Thompson-Ramo-Wooldridge Prods. Co., Los Angeles, Calif.*

"New Magnetic Recordlng Techniques for Data Processing," *M. E. Anderson, Armour Res. Found.*

"Aspects of Magnetic Recording Useful for Industrial Control," *E. G. Wildanger, Man. Applications Engrg., Ampex Corp., Redwood City, Calif.*

Eleventh Annual SWIRECO

April 16–18, 1959, Dallas Memorial Auditorium and The Baker Hotel, Dallas, Tex.

The Eleventh Annual Southwestern IRE Conference and Electronics Show is to be held April 16, 17, 18 at Dallas, under Sponsorship of the Dallas section.

In addition to complimentary breakfasts and luncheons for authors and session moderators, a succession of special events is planned. An opening day luncheon on April 16 will feature an address on "Manpower of Scientific Enginering" by National IRE President, Dr. Ernst Weber. Professor Charles E. Harp, Region 6 Director, Gordon K. Teal, National Board Member, and other IRE officials will join in welcoming industry leaders from the Dallas-Fort Worth area for this occasion.

A reception for Region 6 section officers and conference officials will be held at 5 P.M. opening day, following the afternoon technical meetings.

On Friday morning, April 17, Region 6 officials will convene for breakfast and the annual regional meeting. That noon, SWIRECO board members will gather for luncheon and the annual conference business meeting.

The conference banquet and dance will be held on Friday night preceded by a cocktail hour beginning at 6:30.

Student activities will get underway with a breakfast for school representatives and authors of papers (one from each major school in the region) entered in the student competition. Paper competition will be conducted Thursday afternoon. The three winning papers will be presented in a regular conference technical session on Friday afternoon, and awards will be made at the Friday night conference banquet.

Paralleling the conference technical program and exhibits on Saturday morning will be activities for radio amateurs. Radio amateur day will feature morning and afternoon programs, plus operating demonstrations of latest amateur equipment.

Technical sessions will be held Thursday afternoon, April 16; Friday morning and afternoon, April 17; and Saturday morning, April 18. Site for presentation of papers will be the mezzanine area of the Dallas memorial auditorium, a portion of the same plant which will house conference exhibits.

Thursday Afternoon, April 16

Detection Systems and Techniques

Moderator, *E. L. Hixon, Chairman San Antonio-Austin Section.*

Review paper: "Fundamental Problems in Detection Systems," *C. R. Rutherford, Temco Aircraft Corp., Dallas, Tex.*

"On Measuring Rapidly Fluctuating Radar Echoes," *S. E. Smith, Defense Res. Labs., Austin, Tex.*

"The Signal to Noise Ratio in Infrared Detection Systems," *W. S. Spracklen, Texas Instruments Inc., Dallas, Tex.*

"A Nomograph and Tables of Values, for Use in Making Computations of Maximum Infrared Detection Range," *C. S., Williams, Texas Instruments Inc., Dallas, Texas.*

"Electronic Location of Pipeline Leaks," *F. V. Long, Texas Eastern Transmission Corp., Shreveport, La.*

Communications

Moderator, *J. I. Koski, Chairman, Ft. Worth Section.*

"An Underwater Communications System," *N. D. Miller, Texas Instruments Inc., Dallas, Tex.*

"Tropospheric Scatter, A Modern Communications Technique," *R. M. Mitchell and H. D. Hern, Collins Radio Co., Cedar Rapids, Iowa.*

"Forecasting Television Service Fields," *Dr. A. H. LaGrone, Assoc. Prof. of Elec. Engng., Univ. of Texas, Austin, Tex.*

"Ultra-Simplification of Television Camera Design," *S. M. Zimmerman, The Electron Corp., Dallas, Tex.*

Friday Morning, April 17

Solid-State Electronics

Moderator, *C. Houston, Chairman, Lubbock Section.*

Review paper: "Solid State Devices and Phenomena," *Dr. J. R. Macdonald, Texas Instruments Inc., Dallas, Tex.*

"Thermal Impedance Measurements of Silicon Diodes and Rectifiers," *H. T. Fristoe, Prof. of Elec. Engng., Oklahoma State Univ., Stillwater, Okla.*

"A Survey of Maser Developments," *Dr. C. F. Davis, Texas Instruments Inc., Dallas, Tex.*

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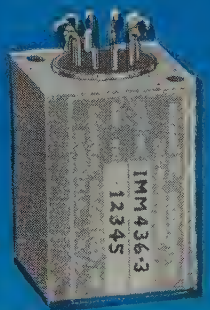

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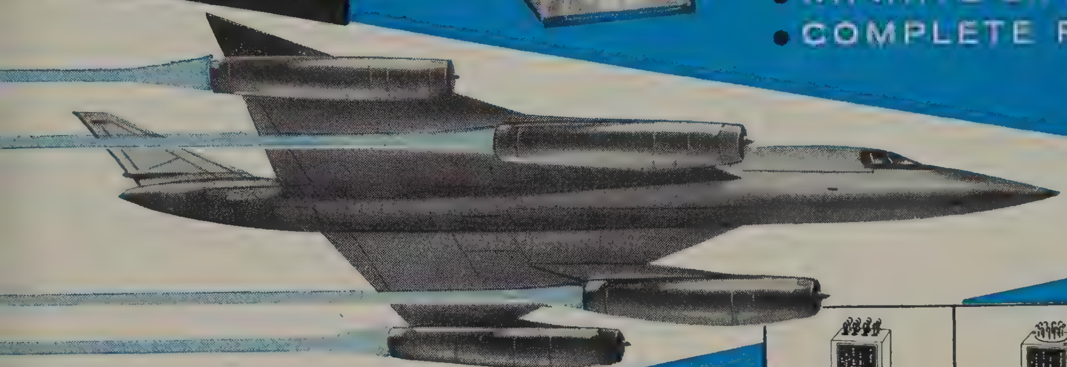

TYPE NUMBER	Magnetic Input Modulator IMM-436-2	Magnetic Input Modulator IMM-436-3	Magnetic Thermocouple Converter MTC-435-2
Excitations Frequency—Carrier	400 cps	400 cps	400 cps
Signal Winding DC Resistance	1000 ohms ±15% each signal winding	1000 ohms ±15% each signal winding	10 ohms ±15%
AC Excitation Volts	5.5 V. @ 400 cps	2.5 V. @ 400 cps	6 V. RMS
Input DC Signal Range	0 to ±100 μa.	0 to ±80 μa.	0 to ±10 mv.
AC Output Range	0 to 2.2V. @ 400 cps (sine wave)	0 to 1.5V. @ 400 cps (sine wave)	0 to 2.7V. @ 400 cps (sine wave)
Overall Dimensions (Inches)	27/32x27/32x1 5/16	27/32x27/32x1 3/16	1 1/4x7/8x5/8
Null Amplitude (Noise Level)	20 mv. RMS	15 mv. RMS max.	25 mv. RMS max.
Output Impedance	7000 ohms	7000 ohms	10,000 ohms
Null Drift (In terms of input signal) −65°C to +100°C	±0.5 μa. max.	±0.5 μa. max.	±0.1 mv. max.
Hysteresis — % of maximum input signal	0.5% maximum	0.5% maximum	0.5% maximum
Type of Mounting	Male Stud	Female Insert	Male Stud
Maximum % Distortion in Output	25%	15 %	20%
Weight Ounces	1.3 oz.	1.2 oz.	1.5 oz.

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"Magnetoabsorption," *W. L. Rollwitz, Southwest Res. Inst., San Antonio, Tex.*

"A Semiconductor Diode Parametric Amplifier at Microwave Frequency," *K. Kotzebue, Texas Instr. Inc., Dallas, Tex.*

Computing Techniques

Moderator, *T. G. Powers, Chairman, Beaumont-Pt. Arthur Section.*

"Computing Fourier Coefficients Without the Use of Multipliers," *A. R. Teasdale, Jr., Temco Aircraft Corp., Garland, Tex.*

"Cut-Product Approximants for Time Delay in Electronic Analog Computers," *J. N. Warfield, Assoc. Prof. of Elec. Engng., Univ. of Kansas, Lawrence, Kan., and L. E. Weaver, Assoc. Prof. of Elec. Engng., Univ. of Arizona, Tucson, Ariz.*

"A Photoformer Function Generator of Two Independent Variables," *L. G. Polimerou, Chance Vought Aircraft, Inc., Dallas, Texas.*

"The Cycle Splitter—A Wide-Band Precision Frequency Multiplier," *B. E. Keiser, Missouri Res. Labs., Inc., St. Louis, Mo.*

New Component Technology

Moderator, *J. D. Reid, Chairman, Little Rock Section.*

"Circuit Fabrication with Thin Films in Microcircuitry," *Dr. C. K. Hager and D. W. McKenzie, Varo Mfg. Co., Garland, Texas.*

"Low Frequency Solion Amplifier," *G. T. Kemp, Texas Res. Assoc., Austin, Tex.*

"An Acoustic Definition Meter Employing Solion Integrators," *C. D. Anderson, Texas Res. Assoc., Austin, Tex.*

"Use of Solions in Automation," *J. J. Moore, Texas Res. Assoc., Austin, Tex.*

Friday Afternoon

Applications of Solid-State Techniques

Moderator, *R. W. Bains, Chairman, Shreveport Section.*

"Very Low Level Transistor Operation," *J. L. Collins and B. F. Weiss, Defense Res. Lab., Univ. of Texas, Austin, Tex.*

"Studies of Surface-Conductance Modulation in High-Resistivity P-Type Silicon," *Dr. C. G. Peattie and Dr. W. R. Savage, Texas Instruments Inc., Dallas, Tex.*

"An Ultrashort Lifetime Apparatus," *W. J. Odom, Texas Instruments Inc., Dallas, Texas.*

"Low Level Transistorized Chopper," *W. Lawrence, Texas Instruments Inc., Dallas, Tex.*

Forum on Science Education

Moderator, *Dr. A. Straiton, Chairman, Dept. of Elec. Engng., Univ. of Texas.*

This session will open with talks by each of four panelists, two presently in industry but who formerly taught in universities and two who are presently in academic circles. Following the talks, a panel discussion will be held. The latter part of the discussion will be devoted to questions from the audience. Panel members are: *Dr. C. F. Squire, The Rice Inst., Dept. of Physics; O. Becklund, Texas Instruments Inc., Dallas, Tex., Dr. F. W. Tatum, Southern Methodist Univ., Dept. of Elec. Engng.; T. A. Wright, Jr., and A. Earl Cullum, Jr., Consulting Engineers, Dallas, Tex.*

Medical Electronics and Winning Student Papers

Moderator, *F. C. Smith, Chairman, Houston Section.*

"The Application of the Ultraviolet Flying-Spot Scanner to Biological Problems," *P. O'B. Montgomery, M.D., Southwestern Med. School, Univ. of Texas, Dallas, Tex.*

"Design and Operation of the Ultraviolet Flying Spot Scanner," *W. A. Bonner, Southwestern Med. School, Univ. of Texas, Dallas, Tex.*

Presentation of third prize student paper.

Presentation of second prize student paper.

Presentation of first prize student paper.

Saturday Morning, April 18

Circuit Theory and Applications

Moderator, *R. L. Atchison, Chairman, Tulsa Section.*

"The RC Triple Parallel-T Network," *G. White, White Instrument Labs., Austin, Texas.*

"An Adaptive Frequency Domain Filter for Missile and Satellite Tracking," *B. E. Keiser, Missouri Res. Labs. Inc., St. Louis.*

"The Use of Thermionic Diodes as Amplifiers, Frequency Doublers, Los Pass Filters, Integrators, True RMS Detectors, Variable Resistors and Choppers," *W. H. Hartwig, Assoc. Prof. of Elec. Engng., Univ. of Texas, Austin, Tex.*

"A Precision Instrumentation System for Minimizing Inter-Channel Time Displacements in Multi-Channel Tape Recorders," *T. E. Owens and R. H. Wallace, Defense Res. Lab., Univ. of Texas, Austin, Tex.*

Miscellaneous Papers

Moderator, *T. McDonald, Vice-Chairman, Oklahoma City Section.*

"Lunar Probe Tracking System," *D. O. Martin and P. M. Green, Collins Radio Co., Dallas, Tex.*

"Scan Conversion," *A. Kuehne, Civil Aeronautics Administration, Oklahoma City, Okla.*

"A Bright Display Indicator for Commercial Airborne Weather Radar," *K. L. Scott, Collins Radio Co., Dallas, Tex.*

"Factors Influencing Sensitivity of Vidicon TV Camera Tubes," *H. Albertine, Jr., General Electrodynamics Corp., Garland, Texas.*

"The Specification of Reliability in Contracts," *J. L. Burnside and S. W. Malasky, Lockheed Aircraft Corp., Sunnyvale, Calif.*

1959 Third National Conference on Analog and Digital Instrumentation

APRIL 20–21, 1959, BELLEVUE-STRATFORD HOTEL, PHILADELPHIA, PENNSYLVANIA

The Third National Conference on Analog and Digital Instrumentation, sponsored by the Recording and Controlling Instrumentation Committee of the American Institute of Electrical Engineers in cooperation with the American Society of Mechanical Engineers-IRD, the Professional Groups on Instrumentation and on Industrial Electronics of the IRE, and the Instrument Society of America, will be held in Philadelphia, Pa. on April 20 and 21.

The purpose of the Conference is to provide a meeting ground for the maker and user of electrical instruments where there can be a mutual interchange of practical technical information on a formal and informal basis, thereby leading to a better utilization of existing instrumentation and to a better direction in the development of new electrical instruments.

There will be luncheons on both days of the conference and a hospitality social hour following the afternoon session on April 20. The Women's Auxiliary of the Philadelphia Section AIEE is arranging appropriate activities for visiting wives.

The following is a program of the seven technical sessions to be held.

Monday Morning, April 20

I. Engineering Education for Instrumentation

Chairman, *L. J. Lunas, Westinghouse Elec. Corp., Newark, N. J.*

"The Status of Electrical Measurements in a Modern Electrical Engineering Curriculum," *D. T. Canfield, Prof. of E. E. at Purdue Univ.*

"Electrical Measurements—in the Core Curriculum," *F. R. Kotter, physicist, Natl. Bur. of Standards, Washington, D. C.*

"Education for Electrical Measurements

in Industry," *S. J. Zammataro, Bell Tel. Labs., Murray Hill, N. J.*

"Control Instrumentation in the Engineering Curriculum," *J. R. Ragazzini, Dean, College of Engineering, New York Univ.*

Formal Discussion on Engineering Education Papers: *Prof. P. A. Borden, Honorary Chairman, University of Dayton; Prof. H. Sohon, Moore School of E.E., Univ. of Pennsylvania;* and *Prof. E. W. Conlon, Director of Research, Drexel Inst. of Tech.*

Monday Afternoon

II. Electronic Analog Recorders

Chairman, *A. J. Williams, Jr., Leeds and Northrup Co., Philadelphia, Pa.*

"A New Concept of Analog Recording," *D. D. Trautner, Minneapolis-Honeywell Regulator Co., Heiland Div., Denver, Colo.*

"A Precision Voltage Reference for Industrial Recorders," *P. B. Robinson, Gen. Elec. Co., Instrument Dept., Lynn, Mass.*

"Evolution of an X-Y Recorder," *F. H. Wyeth and G. S. Talbot, Leeds and Northrup Co., Philadelphia, Pa.*

"X-Y Plotters for Military Service," *F. L. Martinson, Chief Engineer, Electronic Associates, Inc., Long Branch, N. J.*

"Design and Performance Characteristics of a High Speed Wide Chart Recorder," *J. C. Garrigus, the Bristol Co., Waterbury, Conn.*

"A Power Line Transient Recorder," *G. H. Hoshall, Minneapolis-Honeywell Regulator Co., Davies Labs. Div., Beltsville, Md.*

III. Computer Control Systems

Chairman, *R. A. Weiss, Sun Oil Co., Newton Square, Pa.*

"The GE 312 Computer-Controller System," *A. M. Spielberg, Gen. Elec. Co., Computer Section, Tempe, Ariz.*

"Optimizing Control: Theory and Practice," *R. I. Van Nice, Westinghouse Elec. Corp., New Prods. Engineering Dept., Pittsburgh, Pa.*

"A Solid State Digital Computing System for Electrical Load Monitoring," *R. J. Thomas, Union Carbide Nuclear Co., S. Charleston, W. Va.; J. O. Gustafson, Bailey Meter Co., Cleveland, Ohio; and G. E. Foster, Lundell and Co., Chicago, Ill.*

"Digital Techniques in Process Control," *C. E. Jones, Daystrom Systems Div., Daystrom, Inc., La Jolla, Calif.*

Tuesday Morning, April 21

IV. Data Handling Systems for Industrial Processing

Chairman, *A. J. Hornfeck, Bailey Meter Co., Cleveland, Ohio.*

"A Digital Recording and Computing System," *M. S. Blynn, Minneapolis-Honeywell Regulator Co., Brown Instruments Div., Philadelphia, Pa.*

"A General Purpose Computing-Logger for On-Line Process Control Studies," *K. G. Harple and J. Baconnet, Development Engineers, Leeds and Northrup Co., Philadelphia, Pa.*

"A New Stored Program Information System," *G. S. Daniels, Research Manager, Panellit, Inc., Skokie, Ill.*

"The Design of an Industrial Digital Handling System," *J. V. Werme, Data Handling Systems Div., Bailey Meter Co., Cleveland, Ohio.*

"Design of Continuous Industrial Operation Computer Data Handling and Control," *E. J. Otis, Daystrom Systems Div., Daystrom, Inc., La Jolla, Calif.*

V. Electrical Transducer Systems for Electric Control

Chairman, *N. B. Nichols, Taylor Instrument Co., Rochester, N. Y.*

"A Magnetic Amplifier E.M.F. Converter," *H. E. Darling, The Foxboro Co., Foxboro, Mass.*

"Development of an Electronic Differential Pressure Transmitter for Flow Monitoring and Control," *D. J. Aldinger, Taylor Instrument Co., Rochester, N. Y.*

"Electro-Mechanical Transducers for an Electric Control System," *R. Dallimonti* and *J. O. Johnson, Minneapolis-Honeywell Regulator Co., Philadelphia, Pa.*

"Ionization Detector for Vapors," *L. E. Maley, Mine Safety Appliance Co., Pittsburgh, Pa.*

"Electro-Mechanical Balance in Temperature and Pressure Measurement," *E. S. Gilchrist, D. A. Bristol, Manning and Moore, Inc., Danbury, Conn.*

Tuesday Afternoon

VI. High Speed Data Handling and Processing

Chairman, *D. R. Mangold, Gen. Elec. Co., Cincinnati, Ohio.*

"The Beckman Model 210 High Speed Data System," *K. L. Chien, Beckman Systems Div., Anaheim, Calif.*

"The Combi-System—A Proposal for New Concepts in Digital Data Processing," *H. Schwab, Consolidated Electrodynamics Systems Div., Monrovia, Calif.*

"Accuracy Considerations in Analog-to-Digital Converters for Use in High Speed Systems," *R. E. Wright, Epsco, Inc., Boston, Mass.*

"Transistorized Buffer Core Memory for Use with High Speed Simultaneous Measurement of Pressure and Temperature," *S. Reiss, Fischer and Porter Co., Hatboro, Pa.*

"Performance vs Cost Factors in High Speed Data System Specifications," *R. W. Cronshey and S. G. Reque, Gen. Elec. Co., Computer Dept., Phoenix, Arix.*

VII. Electronic Systems for Process Control

Chairman, *E. S. James, E. I. duPont de Nemours and Co., Inc., Louviers Bldg., Newark, Del.*

"Application of Solid State Devices in an Industrial Process Controller," *E. O. Olsen, The Foxboro Co., Foxboro, Mass.*

"Developments in European Controllers," *E. W. Silvertooth, Gen. Precision Equipment Corp., Librascope, Inc., Glendale, Calif.*

"A Digital Valve Actuator," *J. Wapner, Fischer and Porter Co., Hatboro, Pa.*

"Process Control Gas Chromatography," *H. J. Noebels, Beckman Instruments, Inc., Fullerton, Calif.*

"An Integrated Miniature Electronic Control System," *E. J. Cranch and W. B. Alden, Leeds and Northrup Co., Phila., Pa.*

Spring Technical Conference of the IRE Cincinnati Section

April 21–22, 1959, Engineering Society Bldg., Cincinnati, Ohio

Maj. Gen. J. B. Medaris will be the keynote speaker at the Annual IRE Spring Technical Conference. This year's conference with "Electronic Data Processing" as its theme, will be held April 21–22 in Cincinnati, Ohio.

Papers and exhibits will also be presented during the two-day conference on data processing as related to communications, radar, computers, missile technology, chemical engineering, machine tooling, and nucleonics.

The committee for this year's conference includes H. E. Hancock, Chairman; C. Schneider, Advertising; R. T. Schlemmer, Banquet; J. R. Ebbeler, Papers; C. H. Osterbrock, Registration; and J. I. Mika, Publicity.

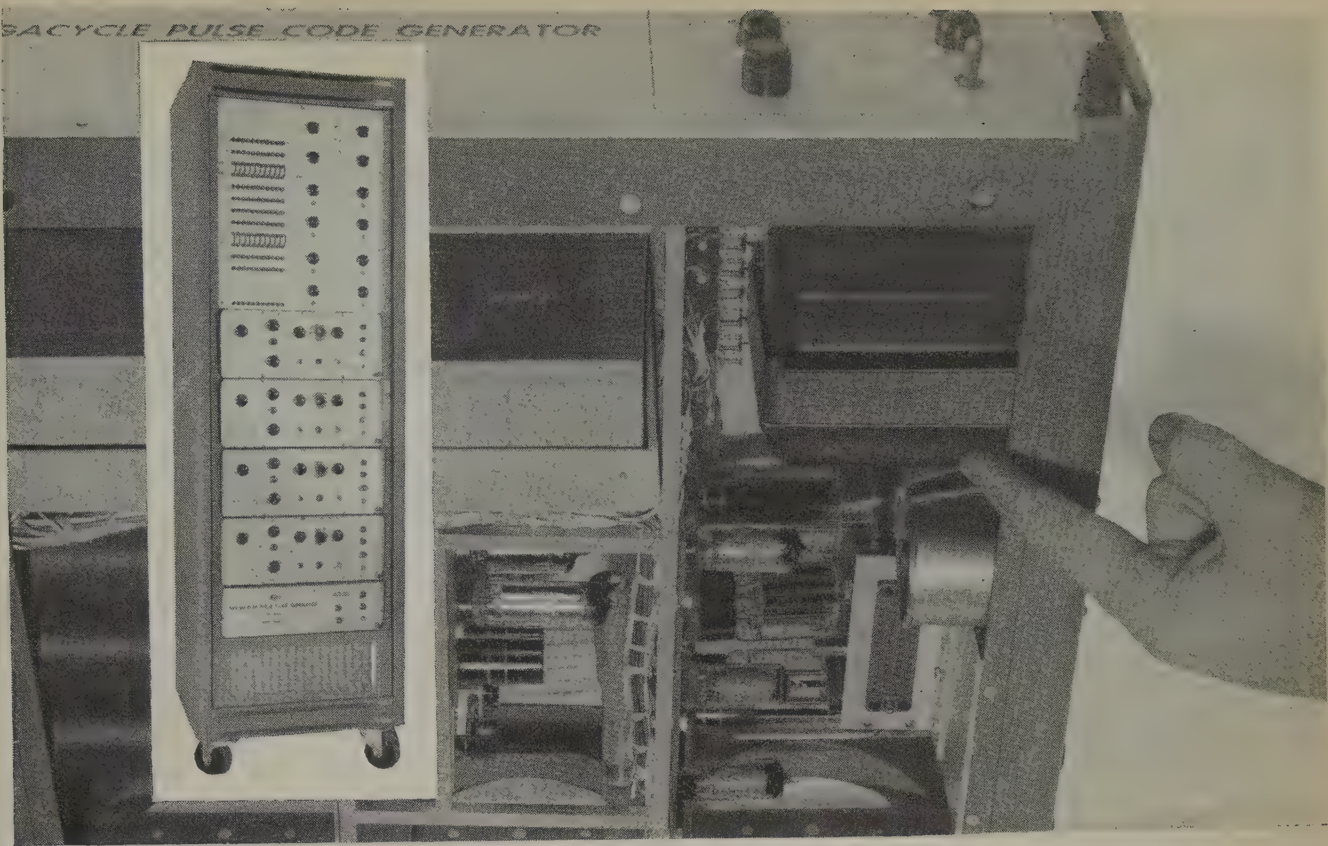
For information regarding registration, contact C. H. Osterbrock at the University of Cincinnati.

The agenda for the Technical Conference is as follows.

Tuesday Morning, April 21

Moderator, *A. B. Bereskin, Prof. of Elec. Engng., Univ. of Cincinnati.*

"Reliability of Very-High-Speed Switching Transistors," *L. Hermansen, Sprague Elec. Co., N. Adams, Mass.*



Sola Constant Voltage Filament Transformer is an integral part of this Electro-Pulse, Inc. Megacycle Pulse Code Generator. It provides regulated filament voltage for reliable operation of the equipment and for full life of its electron tubes.

Sola transformer regulates filament voltage within ±1%--protects tubes from inrush currents and line transients

Fluctuations in supply voltage for electron tube filaments can be costly . . . in shortened tube life . . . in substandard performance . . . in equipment downtime. Electro-Pulse, Inc. solved its filament voltage problems through this straightforward approach: the company's Megacycle Pulse Code Generator includes a Sola Constant Voltage Filament Transformer built-in as part of its power supply.

This versatile unit does the step-down job of a conventional transformer and it also regulates the filament supply — a task that ordinary filament transformers don't pretend to do. Filament voltages are stabilized to within ±1% even with line voltage variations as great as ±15%. Its current-limiting characteristic protects tubes from cold inrush currents upon starting—as well as from line transients. It is a simple, reliable static-magnetic regulator with automatic and virtually instantaneous action. Variations in input voltage are usually corrected within 1.5 cycles. There are no tubes or moving parts, and no manual adjustment or maintenance is necessary.

The filament voltage regulator illustrated is only one of a complete line of Sola Constant Voltage Transformers having wide application in electrical and electronic devices. They include such special types as harmonic-free, plate-filament, and adjustable output units—all provide the benefits of regulated input voltage. More than 40 ratings of these compact, economical regulators are available from stock, and Sola manufactures custom-designed units (in production quantities) to meet special needs.

For complete data write for Bulletin 1D—CVF-269

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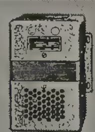
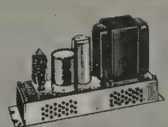


CONSTANT VOLTAGE TRANSFORMERS



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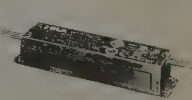
MERCURY LAMP TRANSFORMERS



FLUORESCENT LAMP BALLASTS

A DIVISION OF BASIC PRODUCTS CORPORATION

"High Speed Data Signaling System," *E. A. Irland, Bell Tel. Labs., N. Y., N. Y.*

"Data Processing Aspects of Direct Distance Dialing and Automatic Toll Ticketing Equipment," *J. D. Confield, Stromberg Carlson Co., Rochester, N. Y.*

"A Solid State Data Handling Device," *R. C. Baron and T. P. Bothwell, Epsco, Inc., Boston, Mass.*

Tuesday Afternoon

Moderator, *Dr. E. A. Gamble, Manager, Controls Section, FPL Dept., AGT Div., Gen. Elec. Co.*

"Controlling a Computer from a Remote Substation," *J. C. Richter, Gen. Elec. Co., Evendale, Ohio.*

"The Role of Electronic Data Processing at the Atlantic Missile Range," *B. U. Glass, RCA Data Reduction Ctr., Cape Canaveral, Fla.*

"Data Processing Aspects and Applications of Storage Tubes," *A. S. Luftman, Raytheon Mfg. Corp., Waltham, Mass.*

"A New Method of Real-Time Large-Screen Data Display," *C. L. Ellis, Gen. Elec. Co., Syracuse, N. Y.*

Wednesday Morning, April 22

Moderator, *Dr. C. L. Kober, Vice-President, Weapon Systems Engng., Avco Mfg., Corp., Crosley Div.*

"Design of a General Purpose Data Storage System with Specific Application to Air Traffic Control," *E. Gibson and E. Knoer, Avion Div., ACF Industries, Alexandria, Va.*

"High Speed Arithmetic as it will be Done in the New Generation of Computing Equipment," *R. E. Merwin, IBM Corp., Poughkeepsie, N. Y.*

"New Univac Solid State Computer," *J. R. Popps, Remington Rand-Univac Div., Sperry-Rand Corp., Minneapolis, Minn.*

"Digital Simulation of an Air Traffic Control System," *J. E. Bybee and J. T. Harvey, Crosley Div., Avco Mfg. Corp., Cincinnati, Ohio.*

Wednesday Afternoon

Moderator, *J. F. Jordan, Vice-President, Manufacturing, The Baldwin Co.*

"Magnetic File Drum," *R. Evans, Lab. for Electronics, Boston, Mass.*

"An Automatically-Controlled Vertical Turret Lathe," *A. O. Fitzner, Giddings and Lewis Machine Tool Co., Fond du Lac, Wis.*

"Application of Analog-to-Digital Encoders to Military Electronics and Machine Tool Systems," *C. L. Winder, Baldwin Piano Co., Cincinnati, Ohio.*

"The Usage of Digital Techniques to Effect Analog Control," *R. C. Whiting, Bendix Computer Div., Bendix Aviation Corp., Chicago, Ill.*

1959 Electronic Components Conference

MAY 6–8, 1959, THE BENJAMIN FRANKLIN HOTEL, PHILADELPHIA, PA.

Wednesday Morning, May 6

I. Keynote Address

"New Directions in Component Development," *J. A. Morton, Vice-President, Bell Telephone Labs.*

II. High Speed Data Processing

Moderator, *F. E. Wenger, Air Research and Dev. Command.*

"Functional Components," *R. J. Cypser, IBM Corp.*

"Electronic Components for Future Computers," *N. M. Abov-Taleb, IBM Corp.*

"Magnetic Domain Switching in Evaporated Magnetic Films," *D. W. Moore, Servomechanisms, Inc.*

"The Fabrication and Properties of Memory Elements Using Electrodeposited Thein Magnetic Films of 82-18 Nickel Iron," *I. W. Wolf, H. W. Katz, G. E. Electronics Lab. and A. E. Brain, Stanford Research Inst.*

Wednesday Afternoon

III. Transmission Devices

Moderator, *R. M. Soria, Amphenol Electronics Corp.*

"Microwave High Speed Switch," *R. F. Lucy, Sylvania Electronic Systems.*

"UHF Ferrite Phase Shift Properties," *R. T. Teragawa, W. D. Fitzgerald and M. Weiner, Chu Associates.*

"Dispersive Distributed Parameter Delay Lines Using Ferrite Materials," *J. W. Brouillette and C. Wellmann, G. E. Electronics Lab.*

"Reciprocal Ferrite Scanned Antenna Feed," *D. Caswell, Caswell Electronics Inc.*

"A Versatile Miniature Switching Capsule," *W. J. Fontana, USASRDL and O. M. Hovgaard, Bell Telephone Labs.*

"An Accurate, Stable Beam-Forming Matrix Utilizing Passive Components," *C. H. Field, Sanders Associates.*

"A Microminiature Ferroelectric Digital Storage System," *R. A. Fotland, Horizons Inc.*

Wednesday Evening

IV. Extreme Environments

Moderator, *E. A. Mroz, Bureau of Ships.*

"Fast Neutron Bombardment of Four-Region Semiconductor Devices," *M. M. Weiss, Bell Telephone Labs.*

"Unconventional Environments: Contemporary and Future Requirements for Electronic Components," *E. R. Pfaff and F. E. Graham, Admiral Corp.*

"Transient Effects of Pulse Radiation on Electronic Parts," *L. Long and H. J. Degenhart, USASRDL.*

"Foil Type Solid Electrolytic Capacitor for High Temperature Use," *Dr. Masatsugu Kobayashi, Nippon Electronic Co. Ltd., Japan.*

"Advance Component Design at Various Temperatures to 600°C," *L. F. Kilham and R. R. Ursch, Raytheon Mfg. Co.*

"The Growth of Metal Whiskers on Electrical Components," *S. M. Arnold, Bell Telephone Labs.*

Thursday Morning, May 7

V. Space Electronics

Moderator, *H. K. Ziegler, U. S. Army Signal R and D Labs.*

"Guided Missile Component Reliability in the United Kingdom," *N. B. Griffin, Royal Radar Est., Great Malvern, England.*

"Reliability of Plugs and Sockets in Rockets and Space Vehicles," *I. A. Davidson and R. T. Lovelock, Belling & Lee Ltd., Enfield, England.*

"Silicon Photovoltaic Cells for Space Vehicles," *H. Nash, International Rectifier Corp.*

"A Cadmium Sulfide Solar Energy Generator for Space Vehicles," *D. A. Hammon and F. A. Shirland, The Harshaw, Chem. Co.*

"Frequency Control Devices for Space Applications," *G. M. R. Winkler, USASRDL.*

"A Piezoresistive Accelerometer," *L. E. Hollander, T. J. Diesel and T. A. Perls, Lockheed Aircraft Corp.*

"Requirements for Components for Satellite and Space Vehicles," *H. B. Vallely, U. S. Army Ordnance Missile Command.*

Thursday Afternoon

VI. Electronic Materials

Moderator, *Sidney J. Stein, International Resistance Company.*

"Some Properties of Tantalum Related to Performance of Tantalum Solid Electrolytic Capacitors," *C. C. Houtz and S. Karlik, Bell Telephone Labs.*

"A New 'Square Loop' Ferrite and Its Potential Fields of Application," *B. R. Budny, Allen-Bradley Co.*

"High-Speed Ferrite Core for a Coincident Current Memory," *M. H. Cook, E. H. Melan and E. C. Schuenzel, IBM Corp.*

"New Inductor Ferrites," *H. Lessoff, W. Croft and J. McCusker, RCA, Needham.*

"High Q Ferrites for Frequencies from 2 to 200 Mc/s," *G. G. Palmer, G. E. Electronics Lab.*

"Electrical Characteristics of High-Density, High-Purity Titanate Ceramics," *D. A. Lupfer, G. E. Electronics Lab.*

"High Pressure-High Temperature Research and New Electronic Materials," *A. A. Giardini, USASRDL.*

Friday Morning, May 8

VII. Microminiaturization

Moderator, *A. W. Rogers, U. S. Army Signal R and D Labs.*

"A Miniaturized Ceramic Transformer I.F. Amplifier," *C. M. Stearns, G. E. Electronics Lab.*

"Piezoelectric Ceramic I.F. Filters," *D. R. Curran and W. J. Gerber, Clevite Corp.*

"Plated Component Connections for Micro-Miniature Circuits," *E. A. Guditz, Lincoln Labs., M.I.T.*

"Micro-Module Structural Design," *J. Knoll and H. S. Dordick, RCA, Camden.*

"Transistors and Diodes for Micro-Module Application," *L. Schork, RCA, Camden, N. J.*

"Micro-Miniature High Temperature Stacked Electronic Components," "Part I—Stacked Circuit Building Blocks," *J. E. Beggs and W. Grattidge, G. E. Research Lab.* "Part II—Weight-Volume Advantages in Systems Applications," *P. J. Molenda, G. E. General Engineering Lab.* "Distributed Parameter Networks for Circuit Miniaturization," *C. K. Hager, Varo Mfg. Co. Inc.*

Friday Afternoon

VIII. Semiconductors

Moderator, *D. B. Peck, Sprague Elec. Co.*

"Junction Hall Generator Devices," *A. G. Milnes and E. V. Weber, Carnegie Inst. of Tech.*

"High Gain Silicon Transistors," *H. W. Henkels and T. P. Nowalk, Westinghouse Elec. Corp.*

"A 2000 Volt Silicon Rectifier Diode," *N. J. Herbert, Bell Telephone Labs.*

"Heat Loss and Series Voltage Distribution in the Application of High Voltage, High Current Silicon Power Rectifiers," *G. Mahaffy, C. L. Wallace and F. W. Parrish, International Rectifier Corp.*

"A Review of Parametric Diode Research," *G. C. Messenger, Philco Corp.*

"Fast Variable Junction Capacitors," *C. J. Spector, Bell Telephone Labs.*

"A Novel Method of Fabricating Ceramic Stems," *A. J. Stoeckert, RCA, Harrison.*

1959 Seventh Region Conference and Electronics Exhibit

MAY 6–8, UNIVERSITY OF NEW MEXICO, ALBUQUERQUE, NEW MEXICO

The Albuquerque-Los Alamos Section of the IRE will be host to the 1959 Seventh Region Conference on May 6–8. The Conference theme is "Electronics in Science." All papers were invited. Except for the opening session all sessions and the 175 booths of exhibits will be located in Johnson Gymnasium on the University of New Mexico campus.

The opening session, Wednesday morning, will be held in the ballroom of the Conference Hotel, the Western Skies, and will be followed by the Conference luncheon. Bus service will be provided between the Western Skies and the University of New Mexico.

A field trip (U. S. citizens only), to the Sandia Laboratory Area 3 test facilities, is scheduled for the afternoon of May 7.

The Ladies Activities include a sightseeing tour with luncheon on Thursday to historic Santa Fe with a stop at Santo Domingo Indian Pueblo. The tour will return to the Western Skies in time for the Conference cocktail party, Thursday evening.

Wednesday Morning, May 6

Opening Session

Chairman: *G. A. Fowler, Sandia Lab.*

Welcome Address: *Honorable John Burroughs, Governor of New Mexico.*

Welcome Address: *T. L. Popejoy, President of University of New Mexico.*

Keynote Addresses

"Today's Dilemma in Engineering Education," *G. S. Brown, M.I.T.*

"Problems in Space Research," *F. W. Lehan, Space Electronics Corp.*

"Instrumentation and Man in Space," *T. Freedman, North American Aviation.*

"History of Instrumentation of Nuclear Explosions," *D. B. Schuster, Sandia Lab.*

"Hydrogen Fusion," *J. A. Phillips, Los Alamos Scientific Lab.*

Wednesday Afternoon

The New Electrical Engineering Education

Chairman: *Dean M. E. Farris, Univ. N. Mex.*

"A First Course in Electrical Engineering," *P. R. Clement and W. C. Johnson, Princeton Univ.*

"Physics Courses in the New Arizona Engineering Curriculum," *A. B. Weaver, Univ. of Ariz.*

"A First Course in Electrical Science," *H. G. Booker, N. DeClaris and B. Nichols, Cornell Univ.*

"A Laboratory-Problem-Centered First Course for Electrical Engineering," *F. J. Young and E. M. Williams, Carnegie Inst. of Tech.*

"Special Preparation for the Pre-Graduate Student," *T. F. Jones, Jr., Purdue Univ.*

Instrumentation for Space Research

"Research in the Space Sciences," *R. Jastrow, NASA.*

"Analysis of Radar Echoes Obtained from Earth Satellites 1957 Alpha and Beta," *W. E. Jaye, Stanford Res. Inst.*

"Millstone Hill Radar," *G. Pettingill, Lincoln Lab., Lexington, Mass.*

"Interplanetary Radar," *P. E. Green, Jr. and G. Pettingill, Lincoln Lab.*

Wednesday Evening

Student Paper Competition

Chairman: *A. H. Koschmann, Univ. of New Mexico.*

Instrumentation of Nuclear Explosions

Chairman: *C. B. McCampell, Jr., Sandia Lab.*

"Nuclear Radiation Detection and Its Relation to Modulation Circuitry," *F. B. Brumley, Sandia Lab.*

"High Time Resolution Telemetry by Microwave RF Link," *C. E. Ingersoll, Sandia Lab.*

"Microbarograph Instrumentation," *H. E. Hansen, Sandia Lab.*

"Design and Calibration of a Total Pressure Probe for Dust-Laden Air," *A. R. Kriebel, Stanford Res. Inst.*

Thursday Morning, May 7

Instrumentation for Space Research

Chairman: *(to be announced)*

"Determining Electron Densities in the Ionosphere," *O. C. Haycock and K. D. Baker, Univ. of Utah.*

"Artificial Asteroids as Research Devices," *C. M. Crain, Rand Corp.*

"Design and Performance of Deep Space Tracking and Telemetry System," *R. Stevens and M. H. Brockman, Jet Propulsion Lab.*

"Design and Performance of Satellite and Lunar Payloads," *D. Schneidarman and C. S. Josias, Jet Propulsion Lab.*

"Exploration of the Earth's Radiation Belt," *J. Lindner, Space Tech. Labs.*

Thursday Afternoon

Electronics for Space Medicine

Chairman: *C. S. White, Lovelace Foundation.*

"Electronic Instrumentation on Animals and Humans," *H. Castillo, Holloman AFB.*

"Sensors for Physiological Variables," *W. Welkowitz, Gulton Industries.*

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Constant modulation level
3 volt output into 50 ohms
Low envelope distortion

50kc TO 65MC

New -*hp*- 606A HF Signal Generator

Here at last is a compact, convenient, moderately-priced signal generator providing constant output and constant modulation level plus high output from 50 kc to 65 MC. Tedious, error-producing resetting of output level and percent modulation are eliminated.

Covering the high frequency spectrum, (which includes the 30 and 60 MC radar IF bands) the new *hp* 606A is exceptionally useful in driving bridges, antennas and filters, and measuring gain, selectivity and image rejection of receivers and IF circuits.

Output is constant within ±1 db over the full frequency range, and is adjustable from +20 dbm (3 volts rms) to —110 dbm (0.1 μv rms). No level adjustments are required during operation.

SPECIFICATIONS

Frequency Range: 50 kc to 65 MC in 6 bands.
Frequency Accuracy: Within ±1%.
Frequency Calibrator: Crystal oscillator provides check points at 100 kc and 1 MC intervals accurate within 0.01% from 0° to 50° C.
RF Output Level: Continuously adjustable from 0.1 μv to 3 volts into a 50 ohm resistive load. Calibration is in volts and dbm (0 dbm is 1 milliwatt).
Output Accuracy: Within ±1 db into 50 ohm resistive load.
Frequency Response: Within ±1 db into 50 ohm resistive load over entire frequency range at any output level setting.
Output Impedance: 50 ohms, SWR less than 1.1:1 at 0.3 v and below.
Spurious Harmonic Output: Less than 3%.
Leakage: Negligible; permits sensitivity measurements to 0.1 μv.
Amplitude Modulation: Continuously adjustable from 0 to 100%.
Internal Modulation: 0 to 100% sinusoidal modulation at 400 cps ±5% or 1000 cps ±5%.
Modulation Bandwidth: Dc to 20 kc maximum.
External Modulation: 0 to 100% sinusoidal modulation dc to 20 kc.
Envelope Distortion: Less than 3% envelope distortion from 0 to 70% modulation at output levels of 1 volt or less.
Spurious FM: 0.0025% or 100 cps, whichever is greater, at an output of 1 v or less and 30% amplitude modulation.
Spurious AM: Hum and noise sidebands are 70 db below carrier.
Price: (cabinet) $1,200.00. (rack mount) $1,185.00.

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"Airborne Galvanic Skin Resistance Studies, Preliminary Report," *G. J. D. Shock, USAF Aeromedicine Lab., Holloman AFB.*

"Instrumentation for a Controlled Soft Landing on the Moon," *D. Freebairn, Jr., McDonnell Aircraft Corp.*

The New Electrical Engineering Education

Chairman: *Dean T. L. Martin, Jr., University of Arizona, Tucson.*

"Broadening the Foundation of Machinery Courses," *J. G. Truxal, Polytechnic Inst. of Brooklyn.*

"A Realistic Program in Energy Processing," *R. D. Chanoweth, Case Inst. of Tech.*

"Energy Conversion and Control at Berkeley," *R. M. Saunders, H. C. Bourne, E. C. Guilford, Univ. of California.*

"An Engineering Analysis Course for Superior Students," *G. B. Hoadley, North Carolina State College.*

"A Combined Machinery and Control Systems Laboratory," *L. W. Von Tersch, Michigan State Univ.*

Friday Morning, May 8

Electronics and Thermonuclear Research

Chairman: *J. A. Phillips, Los Alamos Scientific Lab.*

"Controlled Thermonuclear Research at the Lawrence Radiation Laboratory, Livermore and Berkeley," *H. P. Furth, Univ. of Calif. Lawrence Radiation Lab.*

"The DCX Experiment," *R. A. Dandl, Oak Ridge National Lab.*

"Observations on a Deuterium Plasma Compressed by an Axial Magnetic Field," *K. Boyer, W. C. Elmore, E. M. Little and W. E. Quinn, Los Alamos Scientific Lab.*

"High Current Discharges in Thermonuclear Research," *A. C. Kolb, USN Res. Lab.*

"Some Experimental Aspects of the Stellarator," *J. Gorman, James Forrestal Research Center, Princeton Univ.*

The New Electrical Engineering Education

Chairman: *Dean M. A. Thomas, New Mexico State Univ.*

"Circuit Analysis in a Unified Curriculum," *G. F. Paskusz and B. Bussell, Univ. of California at Los Angeles.*

"A General Course in Traveling Waves," *R. K. Moore, University of New Mexico.*

"The Role of an A-C Network Calculator in the Electrical Engineering Curriculum at the Pennsylvania State University," *P. E. Shields, Pennsylvania State Univ.*

"Qualitative Semiconductor Theory," *D. S. Gage, Northwestern Univ.*

"Feedback in Electrical Engineering Curriculum," *G. Franklin, Stanford Univ.*

Electronics for Space Medicine

Papers to be announced.

Friday Afternoon

Instrumentation of Nuclear Explosions

Papers to be announced. Chairman: *T. B. Cook, Jr., Sandia Lab.*

Electronics and Thermonuclear Research

Chairman: *F. Ribe, Los Alamos Scientific Lab.*

"Exploding Wires for a Megajoule Capacitor Bank Protective Fuse," *F. W. Neilson, Sandia Lab.*

"Application of Microwave Diagnostics to Controlled Fusion Research," *D. M. Slager, Los Alamos Scientific Lab.*

"Electronic Engineering Design Problems in Fusion Research," *V. Smith, Univ. of Calif. Lawrence Radiation Lab.*

"The Application of Ignitrons as Switching Devices in Controlled Fusion Research," *E. L. Kemp, G. P. Boicourt and H. K. Jennings, Los Alamos Scientific Lab.*

"Design Considerations for a Thermonuclear Power Plant," *W. B. Myers, Univ. of Calif. Lawrence Radiation Lab.*

National Aeronautical Electronics Conference

MAY 4–6, 1959, BILTMORE AND MIAMI-PICK HOTELS, DAYTON, OHIO

The program of the National Aeronautical Electronics Conference this year will consist of both general sessions and panel sessions. The general sessions will be conducted in the same manner as they have been in the past. The panel sessions will be held to relate some of the leading topics to the conference theme, "Electronic Systems in the Space Age." These panel sessions will be conducted under the direction of selected recognized leaders and will be composed of invited papers which have direct application to the panel topic. Each panel session is scheduled in two parts, the presentation of papers in the morning and a panel discussion of the session topic by the authors under the chairmanship of the moderator in the afternoon.

Detailed information on the various papers is not at this time available. The following, however, is a program of the scheduled sessions and a listing of the various panel moderators.

Monday, May 4

Morning: Invited Papers Sessions

Afternoon: Panel Sessions

"System Design, Prediction, Evaluation, and Test," *sponsored by IAS.*

"Electronic Systems for Air," *moderator, Dr. G. L. Haller, Vice-President and Gen. Man., Defense Electronics Div., Gen. Elec. Co.*

"Military Systems Management," *moderator, N. L. Winter, Manager ECM Div., Sperry Gyroscope Co.*

"Electronic Systems and Space Flight," *moderator, Dr. I. Travis, Vice-President of Res. and Dev., Burroughs.*

"World Wide Communications Systems," *moderator, Col. W. S. Heavner, Chief, 456L SSPO, ARDC, Det. #1.*

Monday Afternoon

General Sessions

"Telemetering," *moderator, J. General, Hq. ARDC.*

"Computers," *moderator, Dr. C. Ross, Chief, Office of Data Syst. Dev., NSA.*

"Radar," *moderator, Prof. J. F. Reintjes, Servo-Mechanisms Lab., M.I.T.*

"Antennas," *moderator, Dr. P. Mayes, Prof. Elect. Eng. Antenna, Research Lab., Univ. of Ill.*

"Reliability," *moderator, R. J. Framme, Directorate of Labs., WADC.*

Tuesday Morning, May 5

General Sessions

"Air Safety," *moderator, D. S. King, Chief, System Engng. Div. FAA.*

"Components," *moderator, H. L. Holley, Radioplane Div., Northrup.*

"Thermal Design," *moderator, W. Robinson, Consulting Eng.*

"Navigation," *moderator, V. I. Weihe, General Precision Equip. Corp.*

"Circuits," *moderator, Dr. A. M. Skellett, Director of Research, Tung-Sol Elec. Co.*

Tuesday Afternoon

Forum

Wednesday, May 6

Morning: Invited Paper Sessions

Afternoon: Panel Sessions

"Impact of Electronic Environment on Airborn Weapons," *moderator, R. J. Nordlund, Technical Director, Weapons Guidance Lab., WADC.*

"System Flexibility Through Modulad Design," *moderator, W. Milnick, Aerial Reconnaissance Lab., WADC.*

"Operations Research in Electronic System Design," *moderator Dr. M. Astrachan, Prof. USAFIT.*

Wednesday Morning

General Sessions

"Civil Avionic System Design," *moderator, W. T. Carnes, Jr., Aeronautical Radio, Inc.*

"Extreme Altitude Environments," *sponsored by IAS.*

Wednesday Afternoon

General Sessions

"Maintainability," *moderator, H. D. Voeglen, Man. Reliability Res., Proj. Sec., RCA.*

"Simulators," *moderator, H. H. Giesecke, Director for System Analysis, Federal Aviation Agcy., Bur. of Res. and Dev.*

Industrial Engineering Notes

(Continued from page 6A)

Panel 4 shall study the system proposals referred to it by Panel 1 with particular regard to 1) the performance of existing monophonic receivers when tuned to the stereophonic signal (receiver compatibility), 2) the performance of stereophonic receivers designed for the stereophonic signal (stereo performance) and 3) the performance of stereophonic receivers when tuned to monophonic signals (reverse receiver compatibility). *Panel 5—Field Testing:* Panel 5 shall study and compare the system proposals referred to it by Panel 1 and the existing services with particular regard to coverage, interference effects and other matters related to channel utilization; and shall conduct field tests with the advice and assistance of the other Panels. *Panel 6—Subjective Aspect:* Panel 6 shall provide to the other panels the available scientific information on the subjective aspects of the stereophonic reproduction of sound . . . **Chairmen for the six National Stereophonic Radio Committee panels were announced by Dr. W. R. G. Baker, Director of EIA's Engineering Department.** They are: *Panel 1—System Specifications,* C. J. Hirsch, Executive Vice-President, Hazeltine Res. Corp., chairman. *Panel 2—Interconnecting Facilities,* A. Jensen, consultant and recently retired Director of Audio and Visual Engineering, Bell Tel. Labs., chairman. J. M. Barstow, Bell Tel. Labs., vice-chairman. *Panel 3—Broadcast Transmitters,* R. N. Harmon, Director and Vice-President for Engineering, Westinghouse Broadcasting Co., chairman. *Panel 4—Broadcast Receivers,* J. N. Benjamin, President of David Bogen Co., chairman. *Panel 5—Field Testing,* A. P. Walker, Manager—Engineering Dept., Natl. Assoc. of Broadcasters, chairman, R. H. Beville, Vice-President for Engineering, Radio Stations WWDC and WWDC-FM (Washington, D. C.), vice-chairman, *Panel 6—Subjective Aspects,* Dr. A. N. Goldsmith, Consulting Engineer, Chairman. *Coordination Committee,* D. G. Fink, Philco Corp., chairman, and W. J. Morlock, Gen. Elec. Co., vice-chairman. Chairman of the new NSRC Committee is C. G. Lloyd, of General Electric Co. George H. Brown, of RCA, is vice-Chairman.

Military Electronics

Development of a new, cool-running radio tube, considered the first major breakthrough in basic tube design in more than 30 years, was announced by the Army. The new device that glows blue instead of red and uses less than one-tenth the power of a standard hot cathode tube was developed jointly by the Signal Corps and Tung-Sol Electric, Inc. "It may have as great an impact on electronics as the discovery of the transistor," the Army said. The "cold" cathode tube principle is believed adaptable to almost all types of electron tubes, including TV screens, giant

(Continued on page 36A)

Industrial Engineering Notes

(Continued from page 34A)

radar and transmitting tubes, as well as nearly all general-purpose radio tubes. Miniature and sub-miniature tubes, which would compare favorably with the transistor in size, may also be feasible. . . . **Dr. T. Keith Glennan, National Aeronautics and Space Administrator, said the space program will have available a total of some $830 million in fiscal year 1960 to be spent by the NASA, Advanced Research Projects Agency (ARPA), and the Atomic Energy Commission.** In addition to the ultimate plans for a 1.5 million pound thrust engine and other expenditures on propulsion projects, Dr. Glennan said, "The development of midcourse and terminal guidance systems clearly seems to be necessary. Above all, simplicity and reliability must be built into these systems. Our principal task at the moment is to acquire information about space environment—and this calls for reliability in getting instruments aloft, whenever and wherever we may want to make our measurements. Once we have this family of boosters and propulsion vehicles, reliably guided and controlled, we will be able to concentrate fully on the principal tasks that face us—the acquisition of new knowledge and the undertaking of manned flight in space." . . . **"Proceedings of the Statistical Techniques in Missile Evaluation Symposium" are available upon request at no cost.** Mail requests to: Office of Ordnance Research, U. S. Army, Duke University, Durham, North Carolina.

Governmental and Legislative

The Laboratory Division of the FCC's Office of Chief Engineer released a report (L. D. 6. 3. 1—"Field Strength Measurements"). The report contains much basic practical information of interest to engineers and other technical personnel engaged in making field strength measurements of all types. A copy of the report may be obtained from the Technical Research Division, Room 7506, New Post Office Building, Washington 25, D. C., upon individual request. . . . **A major advance in television and FM radio broadcasting and reception—adapting these two broadcasting systems to stereophonic sound—has been developed by Philco Corp., the company reported.** Philco revealed its systems which cover both TV and FM stereo broadcasting as well as stereo TV receivers and FM stereo radio receivers in two petitions filed last week with the Federal Communications Commission. Philco on Dec. 3, 1958 filed a petition with the FCC covering its AM stereo broadcasting and home radio receivers development (Weekly Report, Vol. 14, No. 43). The AM stereo petition as well as the television and FM radio stereo petitions were filed by Henry B. Weaver, Philco's Washington attorney. The two petitions now being filed ask the FCC to establish an experimental field test pro-

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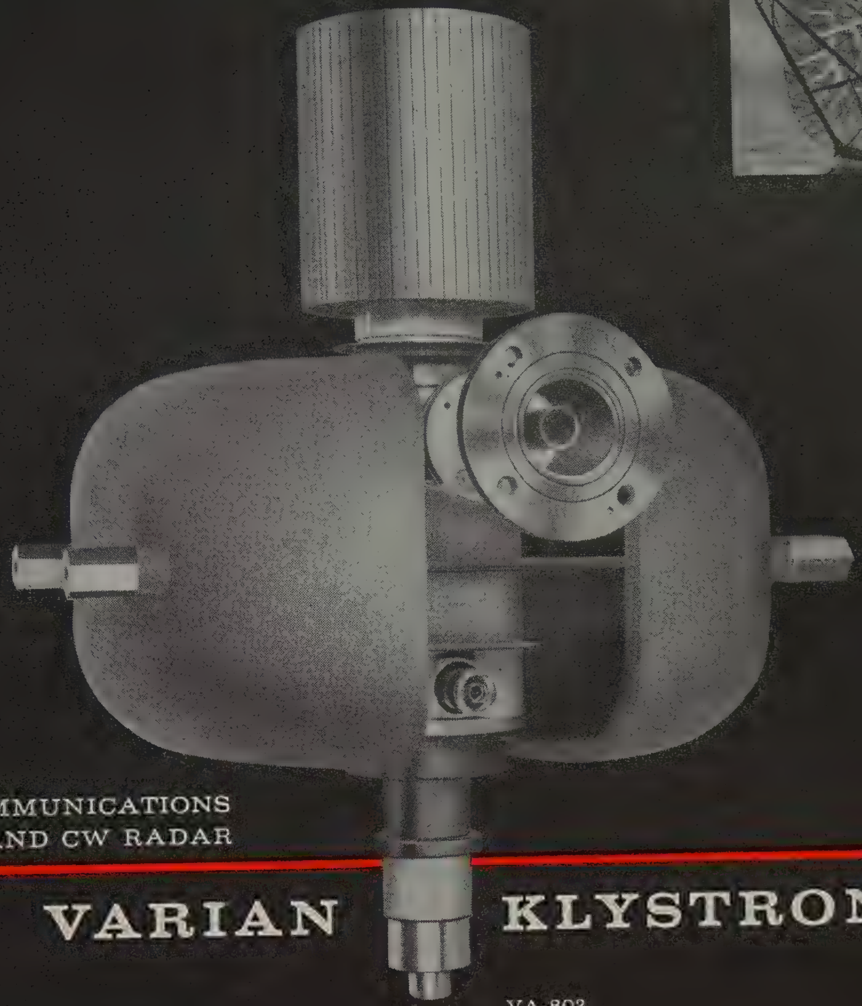

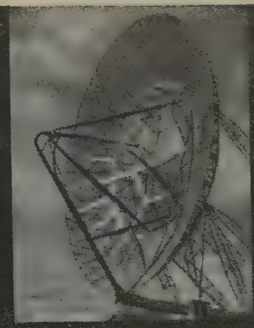

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VA-800	1.7 to 2.4 kMc	10kW cw
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VA-804	4.4 to 5.875 kMc	2kW cw
VA-805	5.875 to 6.425 kMc	2kW cw
VA-806	7.125 to 8.5 kMc	2kW cw
VA-822	9.9 to 10.8 kMc	1kW cw
VA-833A, B	.47 to .985 kMc	10kW cw

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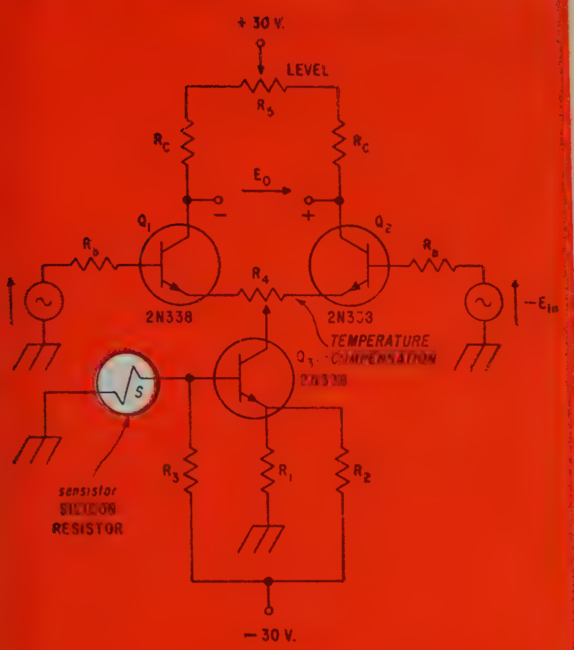
Industrial Engineering Notes

(Continued from page 36A)

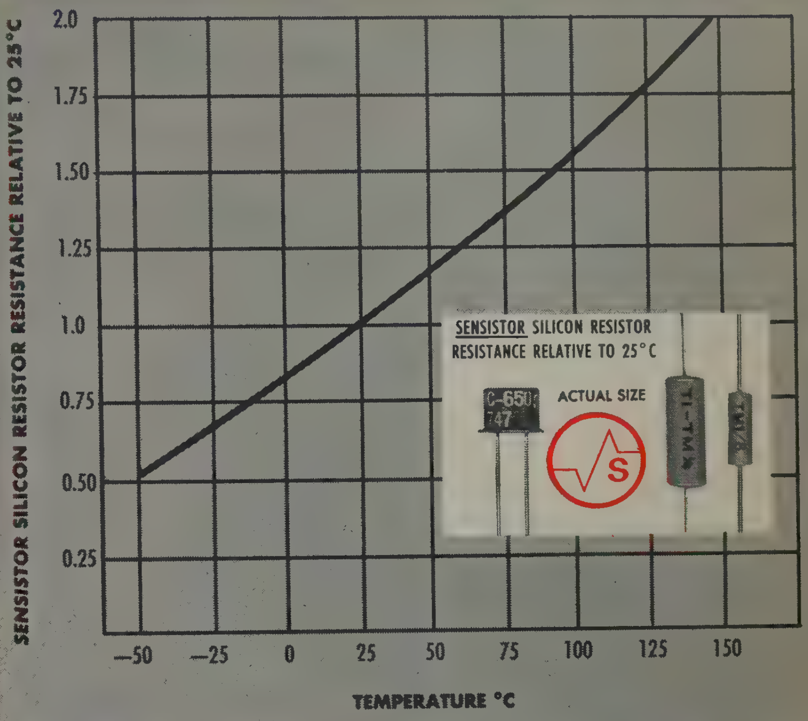
gram for the purpose of testing Philco's system under normal broadcast conditions and upon successful completion of the field test program to establish transmission standards for stereo television and FM radio broadcasting based upon Philco's research and engineering developments. The two FCC petitions are the initial steps toward achieving stereo standards in television and FM broadcasting. Field testing and final FCC approval of standards must follow and could take a year or more. The proposed stereo television and FM radios are easy to tune and reliable in performance. Such receivers will receive stereo broadcasting within the same areas as comparable television and FM monophonic receivers. The Philco proposed system of stereo broadcasting will not reduce the service range of existing transmitters nor will it require any revision of present frequency assignments. In the matter of field testing, Philco has told the FCC that it will cooperate with any licensed FM radio or television broadcasting station, and with the National Stereo Radio Committee as well as the FCC. Philco has also invited the FCC to attend a demonstration of the new stereo systems broadcasting at the Company's Research and Engineering Labs in Philadelphia. The Company now has three stereo broadcasting petitions filed with the FCC, covering television broadcasting and AM and FM radio broadcasting. Highlights of these new broadcasting developments are, no new television or radio frequencies will be needed; no revision of existing television or radio frequency assignments required; all systems fully compatible with single sound broadcasting, and the broadcast stations can switch from monophonic to stereophonic sound at any time. After standards are approved for television and FM radio additional time will be required for broadcasters to prepare for actual television and FM broadcasts and for set manufacturers to build adapters for standard television receivers and FM radio receivers and new stereo receivers for both fields. "Existing receivers would continue to obtain the service now available, but in addition, the new stereo type receivers and existing receivers with stereo adapters would have the benefit of the new service," stated David B. Smith, Vice-President, Technical Planning. "We have completed our development work in this field for television and FM radio and we believe that the matter should be brought to the attention of the broadcast industry generally, and a field test be established," Mr. Smith added. Philco's program has included development of a system which would provide for inexpensive stereo receivers for both television and FM at a price level which everyone could afford and which would put stereo sound reception within the reach of the entire public, it was stated in the FCC petitions.

(Continued on page 40A)

TI APPLICATION NOTE



DIFFERENTIAL DC AMPLIFIER USING *SENSISTOR* SILICON RESISTOR.
Write on your company letterhead for additional applications and data sheet.



HOW TO INCREASE DIFFERENTIAL DC AMPLIFIER STABILITY WITH *sensistor** SILICON RESISTORS

Low drift transistor amplifier circuit using *sensistor* silicon resistor gives drift performance superior to vacuum tube amplifiers for low source impedance applications.

The sensistor silicon resistor has a unique positive temperature coefficient of +0.7%/°C plus a constant rate of change as shown in the graph to the right. Over a 15°C temperature range, the *sensistor* silicon resistor's temperature-resistance curve approaches linearity to an extent that allows its use as a compensating component in a differential D-C amplifier.

This low drift amplifier finds a wide range of low source impedance applications in airborne telemetry where the performance of other types of D-C amplifiers is limited by weight requirements, acceleration, shock, and vibration. It is particularly useful with low level transducers such as thermocouples, strain gages and accelerometers.

DESIGN CONSIDERATIONS

TI 2N338 silicon transistor provides excellent performance as a low drift DC amplifier when used in circuits such as the one shown above.

For optimum performance keep $(2R_B + R_4)$ as small as possible, preferably less than 2000Ω, and the collector currents of Q_1 and Q_2 should remain below 100 μA.

Drift cancellation featured in an uncompensated differential configuration provides an amplifier with an equivalent input drift of 400 μV/°C or less with standard production transistors.

Drifts as low as 6μV/°C will result if the compensating circuit composed of Q_3, sensistor resistor S and their biasing resistors is used with a matched pair of transistors.

CIRCUIT OPERATION

Sensistor resistor S and its biasing resistor R_3 serve as a voltage source which has an output linearly related to temperature . . . level potentiometer R_5 adjusts output voltage E_0 to zero when E_{in} is zero . . . potentiometer R_4 adjusts for minimum output drift due to ambient temperature changes.

As temperature increases, the resistance value of S also increases causing the base of Q_3 to go more negative, thereby reducing the collector current of Q_3. This temperature-dependent current is fed into the differential amplifier through R_4.

Depending on the wiper position of R_4, the correcting signal may be positive, negative or zero. When the wiper is centered, zero correction results. As temperature increases, output voltage E_0 tends to go more positive if the R_4 wiper is placed nearer the Q_2 emitter and negative if the wiper is placed nearer Q_1. The optimum setting for R_4 can be determined by cycling over the desired temperature range to give a minimum drift for changes in ambient temperature.

*TRADEMARK OF TEXAS INSTRUMENTS

from THE WORLD'S LARGEST SEMICONDUCTOR PLANT

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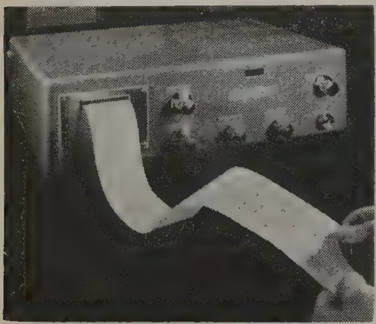
INCORPORATED
SEMICONDUCTOR-COMPONENTS DIVISION
POST OFFICE BOX 312 · 13500 N. CENTRAL EXPRESSWAY
DALLAS, TEXAS

NEWS New Products

Printing Integrator

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

A new instrument which automatically prints out chart area measurements at a rate up to 6,000 counts per minute, has been developed by the **Instrument Div., Perkin-Elmer Corp.,** Norwalk, Conn. Designed for use with self-balancing potentiometer recorders, it replaces previous instruments which merely marked pips, which later had to be totaled by hand.



The Model 194 Printing Integrator is applicable wherever the simultaneous integration of recorder chart readout is needed. Typical uses include: integration of stress-strain curves for the determination of cyclic and rupture energies; continuous process weighing, where the integrator totalizes the amount of material supplied to a conveyor belt; integration of rocket thrust in conjunction with transducers to obtain propulsion efficiency; and, obtaining of daily totals of a process plant flow.

The integrator gives printed numerals on standard adding machine tape, and may be operated automatically or manually.

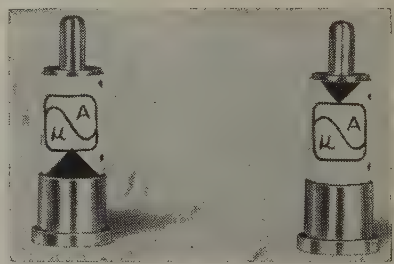
In automatic operation, the new integrator prints digital integrals as curves or peaks appear on a recorder chart. The integrator's tape travels continually at the same rate as the chart paper, so that tape and chart records later may be compared.

The printing mechanism is activated by sensor in a potentiometer assembly installed in the recorder. The sensor energizes the printer whenever the recorder pen begins to move up-scale, after a down-scale movement. Successive integrals are similarly recorded whenever the recorder pen travels through a minimum point of inflection. When the recording is completed, the final integral is printed manually.

To determine relative areas, the area under each segment of the total curve is obtained by subtracting the intregal printed at the beginning of the next segment. Percentage areas are calculated readily by totaling all areas and normalizing. At a chart speed of ½ inch per minute, the integrator produces 1,263 counts per square inch of recorder chart, thus permitting extremely precise measurements.

Varactor Diodes

Microwave Associates, Inc., Burlington, Mass., announces the availability in production quantities of the new silicon varactor. The varactor is a diffused silicon PN junction diode designed to be a variable capacitance with low loss at high frequencies. The unit complies with MIL-E-1 outline 7–1 for cartridge type crystal rectifiers and will fit most standard crystal holders. In the standard MA-460 series, the pin end of the diode is connected to *P*-type material on the top of a small "mesa" and the *N*-side of the silicon element is connected to the base. The reversed polarity denoted by the suffix R is also available. The mechanically reversible MA-450 series may also be ordered with longer delivery time but the single-ended units are generally recommended because they insure placement in holders with the proper end in contact with a heat sink. The minimum cutoff frequencies are graded in 10 mc steps starting with the suffix *A* at 20 kmc. The units are currently available as high as 60 kmc. The varactor is useful in low-noise diode amplifiers (also known as reactance amplifiers or parametric amplifiers), amplifying up-converters, harmonic generators, high-level modulators, frequency dividers, voltage-variable tuners, switches, reactive limiters, and high-speed computers. Low-noise amplifiers have been constructed from 1 mc to 6000 mc. Harmonics have been generated as high as 100 kmc. At lower frequencies, conversion



(Continued on page 135A)

Industrial Engineering Notes

(Continued from page 38A)

Industry Marketing Data

Cumulative television output during the January-December period of last year amounted to 4,920,428 compared with 6,399,345 TV receivers made in the same 12-month period in 1957. Cumulative radio receiver production during the calendar year 1958 totaled 12,577,243 including 3,715,362 automobile receivers, compared with the 15,427,738 radios made during calendar year 1957 which included 5,495,774 automobile receivers. Cumulative receiving tube sales for the calendar year 1958 totaled 397,366,000 valued at $341,929,000 compared with the 456,424,000 tubes worth $384,402,000 sold during calendar year 1957. Cumulative television picture tube sales during the calendar year 1958 totaled 8,252,480 tubes worth $163,482,674 compared with the 9,721,008 TV picture tubes sold in calendar year 1957 with a dollar value of $183,231,337.

Engineering

The National Bureau of Standards, in Tech. Rep. No. 2274, detailed its Telemetering Transducer Program. To summarize, the NBS said: The telemetering of flight data—an essential part of missile and aircraft testing—is often limited by a lack of information on performance of transducers. To obtain the necessary data on these important devices, which convert mechanical quantities to electrical signals, the NBS is carrying on a special program of research, development, and testing. Such studies predict transducer performance under the environmental conditions encountered in flight. Also, in the course of the work, transducers have been calibrated in terms of their dynamic response, making possible the improved choice of transducers for telemetering of such important quantities as pressure and acceleration changes. . . . **"Techniques for Accurate Measurement of Antenna Gain" was issued by the NBS.** The booklet is Circular 598 and is priced at 15 cents each. It can be obtained from the Government Printing Office, Washington 25, D. C. The circular describes techniques developed at the Bureau to minimize the experimental error and thereby increase the reliability of measurements of antenna gain. Special features of the instrumentation, including methods for minimizing and measuring matching losses, are described. . . . **A report of Navy research which showed that ultrasonic welding can be used for continuous-seam-type welds and spot-type solid-state junctions in aluminum structural alloys has been released to industry through the OTS, Commerce Dept., Washington 25, D. C. ("Ultrasonic Welding of Structural Aluminum Alloy," #PB 131680. $2.25 per copy).**

Creative Microwave Technology

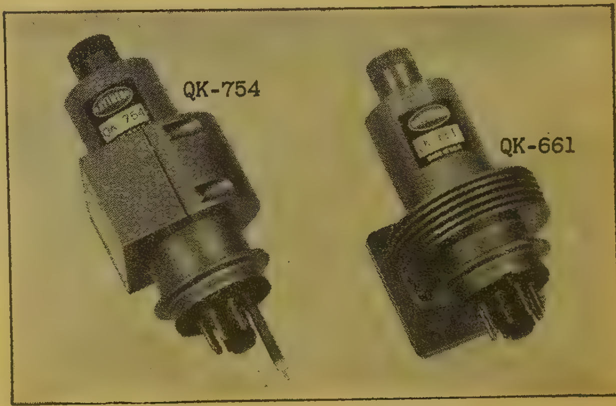
Published by MICROWAVE AND POWER TUBE DIVISION, RAYTHEON MANUFACTURING COMPANY, WALTHAM 54, MASS., Vol. 1, No. 2

NEW ONE-WATT COMMUNICATION KLYSTRONS COVER GOVERNMENT AND COMMON CARRIER BANDS

Designed primarily for use in microwave relay links, the QK-661 and the QK-754, one-watt transmitter klystrons, operate at frequencies of 7,125 to 8,500 Mc and 5,925 to 6,425 Mc, respectively. The QK-661 is the first tube of its kind to cover the entire government band. The QK-754 is the first of a planned series of tubes to cover the entire communications band.

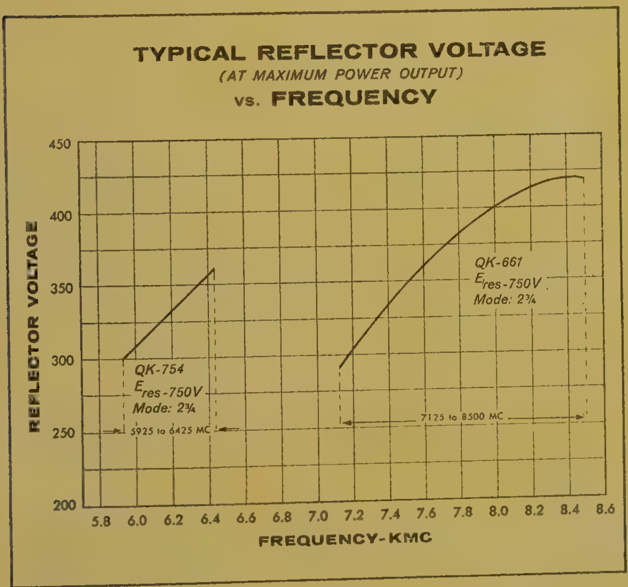
Both are mechanically tuned, integral-cavity, long-life, reflex-type tubes. The QK-754 uses a coaxial output; the QK-661, a waveguide output.

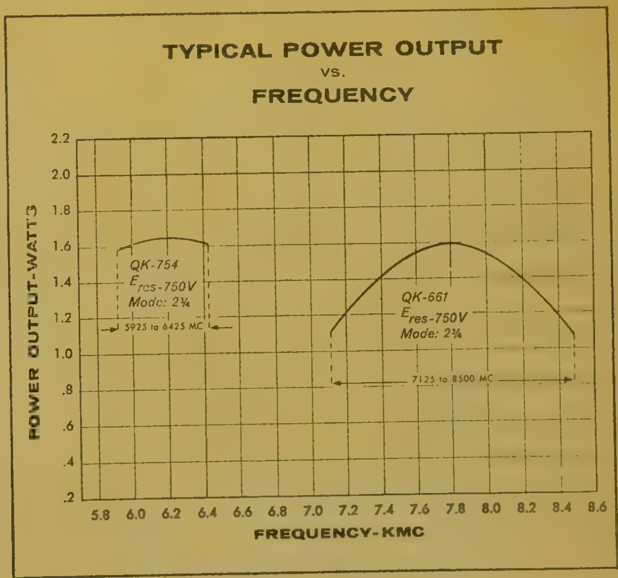
To insure efficient operation the tubes are available with integral cooling fins or with a heat-sink attachment suitable for connection to the chassis.



Typical operating characteristics

	QK-754	QK-661
Frequency Range	5925 to 6425 Mc	7125 to 8500 Mc
Power Output	1.5 watts	1.6 watts
Electronic Tuning (to half-power pts)	50 Mc	25 Mc
Modulation Sensitivity (10 V pk-to-pk mod volt)	1 Mc/V	600 Kc/V
Temp. Coefficient	± 0.1 Mc/°C	± 0.1 Mc/°C





You can obtain detailed application information and special development services by contacting: Microwave and Power Tube Division, Raytheon Manufacturing Company, Waltham 54, Massachusetts

Excellence in Electronics

RAYTHEON

A LEADER IN CREATIVE MICROWAVE TECHNOLOGY

IRE People

Walter J. Albersheim (A'43–SM'48–F'57), formerly of Bell Telephone Labs., has joined the staff of Spencer-Kennedy Labs., Inc., Boston electronics concern, in the capacity of chief engineer. He will head all engineering and development work of the organization, both in the instrument division and in the closed circuit and community antenna television division.

Having received his Doctor of Engineering Degree from Aachen Institute of Technology, Germany, in 1924, Dr. Albersheim came to this country and in 1929 joined the Bell Telephone System as an engineer. From 1941 to January, 1959 he has been on the staff of Bell Telephone Labs. in engineering, research, and supervisory capacities. Dr. Albersheim has been a frequent contributor of papers to technical publications and proceedings of technical societies, and he holds individually or jointly many patents in electronics, both in and out of the field of communications. He is a Fellow of the New York Academy of Sciences and associate member of the Society of Motion Picture and Television Engineers.

❖

The Burroughs Corporation has announced the appointment of **R. W. Anderson** (S'44–A'46–M'55) to the position of manager of quality assurance. Mr. Anderson, who will assume responsibility for the testing, inspection, and quality control of Burroughs 205 and 220 electronic computer systems, is an engineering graduate of Purdue University. He is a member of the American Institute of Electrical Engineers.


R. W. Anderson

❖

Thurb D. Cushing (S'40–A'42–M'51) has been promoted to a new position as chief engineer of Lenkurt Electric Co. of Canada, Ltd. at Burnaby, British Columbia. He has been with Lenkurt since 1957 and was previously applications engineering manager at Lenkurt's U. S. headquarters in San Carlos, Calif. In his new position he will direct engineering of the multichannel carrier and microwave systems and related equipment which the firm manufactures and markets to customers throughout Canada.


T. D. Cushing

Mr. Cushing's experience includes eight years with North-West Telephone Co. in Vancouver where he was responsible for major radio engineering projects, including the British Columbia portion of the Trans-Canada TD-2 system. Earlier positions with Electronic Laboratories of Canada and the National Research Council of the Department of National Defense gave him wide experience in other important radio communications projects.

❖

Beverly Dudley (J'24–A'27–M'43–SM'43), editor of *The Technology Review* at the Massachusetts Institute of Technology since 1945, has resigned to accept a position as assistant to the director of Lincoln Laboratory in Lexington, Mass.

Mr. Dudley, a native of Chicago, studied at the Armour Institute of Technology, now Illinois Institute of Technology, was graduated as an electrical engineer from M.I.T. in 1935, and has the M.S. degree from Columbia University. Before becoming editor of *The Technology Review*, he was associated with the McGraw-Hill Publishing Co. for nine years, successively as assistant and associate editor, managing editor, and western editor of *Electronics*. He was managing editor of *Photo Technique*, consulting editor for McGraw-Hill technical books on radio communication and was the originator of the Radio Communication Series.

Upon graduation from M.I.T. Mr. Dudley became an engineer for RCA Manufacturing Co. and worked on the development of electron tubes. During World War II he was also instructor in electrical communication at Newark College of Engineering. He is the author of "Basis of Radio Communication" in the "Radio Engineering Handbook" (McGraw-Hill) and part author and co-editor of "Handbook of Photography" by Henney and Dudley (McGraw-Hill). He has been editor of an M.I.T. monthly technical bulletin, *Reports on Research*, for ten years.

❖

Philco Corporation has announced the appointment of **Donald G. Fink** (A'35–SM'45–F'47), formerly a director of commercial research, as Director of Research succeeding **David B. Smith** (A'35–SM'44–F'48) who has been named to the president's staff as Vice-President of Technical Planning.

During 1958, Mr. Fink was president of the IRE and in December, 1958 received the first Technological Award of the Year from the New York Institute of


D. G. Fink

(Continued on page 44A)

(Continued from page 42A)

Technology in recognition of his work with the IRE. He is the author of numerous books, including "Color Television Standards," "Engineering Electronics," "Principles of Television Engineering," and "Microwave Radar." During World War II he was a member of the Radiation Laboratory at MIT and headed the Loran division. He served with the Secretary of War as a consultant on radio navigation and radar and participated in the atom bomb tests at Bikini in 1946 as a civilian consultant.

Mr. Fink is a Fellow of the AIEE and the SMPTE and a member of Tau Beta Pi, Sigma Xi and Eta Kappa Nu. In 1951 he was awarded the Radio Fall Meeting Plaque for his contributions to the television industry. Mr. Fink is a graduate of MIT and received his M.S. degree from Columbia University.

After receiving his M.S. degree from MIT in 1934, Mr. Smith joined Philco and held various posts connected with research and engineering until 1938, when he was appointed technical consultant to the vice-president in charge of engineering. He was made director of research in 1941 and in this capacity directed the fundamental microwave and ultra-high frequency research that led to the production of many important types of airborne radar used by the Armed Forces. In 1945 he was made vice-president of research for the corporation. In this newly-created position, Mr. Smith will be responsible for the over-all planning of the corporation's activities in new fields and new product development and will coordinate the planning for all divisions of the Company.

Mr. Smith played a major part in the establishment of standards for television broadcasting as a member of the first National Television System Committee and a vice-chairman of the second National Television System Committee. He is a past IRE section chairman, a director and Fellow of the AIEE. He is a member of Tau Beta Pi, Eta Kappa Nu, Sigma Xi, and the American Association for the Advancement of Science. In 1954 he received the industry's RETMA award for his contribution to the creation of color television standards.

❖

John L. Collyer, chairman of the B. F. Goodrich Company, has announced the election of **Amory Houghton, Jr.** (A'57) as a director of the rubber company. Mr. Houghton is staff vice-president and a director of the Corning Glass Works. He is also a vice-president and director of the Corning Glass Works of Canada, Ltd., and a director of the Dow Corning Corp., the Pittsburgh Corning Corp., and the Sylvania-Corning Nuclear Corp.

❖

(Continued on page 47A)

new transistors from Sprague*

SUPER HIGH-SPEED SWITCHING TRANSISTORS TYPE 2N501

	Typical	Maximum	Units
Rise Time (t_r)	9	18	mμsec
Storage Time (t_s)	9	12	mμsec
Fall Time (t_f)	7	10	mμsec

In circuit with current gain of 10 and voltage turnoff.

Also available as special type 2N501A for 100° C. maximum storage and junction temperatures.

This table tells the story. Sprague Type 2N501 germanium micro-alloy diffused-base transistors are the fastest mass-produced transistors available anywhere! They are unexcelled for high-speed computer applications. The ultra-low rise, storage, and fall time cannot be matched by any other transistor.

Ultra-precise process control in manufacture results in superb and consistent high quality. The basic electrochemical process of fabrication takes the guesswork out of transistor manufacturing. The result is outstanding uniformity of product.

Because of the electrochemical process, Sprague is able to fabricate a graded-base transistor with no intrinsic base region. The Type 2N501 can thus maintain its super high-speed switching characteristics right down to its saturation voltage, providing all the advantages of direct-coupled circuitry with no impairment of switching speeds.

Type 2N501 Transistors are available from Sprague now at extremely reasonable prices. They are transistors you can use today! You need not delay your development work for the future when you design high-speed switching circuits with Type 2N501 Micro-Alloy Diffused-Base Transistors.

Write for complete engineering data sheet to the Technical Literature Section, Sprague Electric Company, 235 Marshall Street, North Adams, Massachusetts.

* *Sprague micro-alloy, micro-alloy diffused-base, and surface barrier transistors are fully licensed under Philco patents. All Sprague and Philco transistors having the same type numbers are manufactured to the same specifications and are fully interchangeable.*

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ELECTRONICS AND COMMUNICATION FOR HOME, INDUSTRY AND DEFENSE

IRE People

(*Continued from page 44A*)

Samuel H. Goldstein (A'57) has been appointed general manager of Epi-Hab L. I. Inc., the first industrial plant operating on the East Coast employing only epileptics. In operation since April, 1958, Epi-Hab has twenty-seven epileptics now employed at its Parsons Boulevard, Jamaica plant and is affiliated with Epi-Hab U. S. A. Training programs are in process at Epi-Hab in soldering, harness wiring, and connector assembly for the electronics industry. A similar plant, Epi-Hab L. A. Inc., has been serving electronic and aircraft companies in Los Angeles, Calif. since 1955.

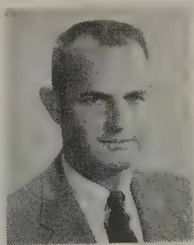
S. H. Goldstein

Prior to joining Epi-Hab, Mr. Goldstein was supervisor of the staff and administrative services section at the W. L. Maxson Corp., in New York, N. Y., handling advertising, publicity and administrative assignments. From 1938–1952 he was with the U. S. government in various administrative capacities.

Mr. Goldstein is publicity chairman of the New York Metropolitan chapter of the IRE Professional Group on Aeronautical and Navigational Electronics and news editor of the PGANE Transactions.

❖

Maurice H. Kebby (S'46–A'48–M'55) has been named chief engineer of the commercial products division of Lenkurt Electric Co. He will direct component and systems development and quality control of Lenkurt telecommunications systems for telephone companies, railroads, and other utilities and government agencies using commercial equipment.



M. H. Kebby

Joining Lenkurt in 1951 as a filter design engineer, Mr. Kebby was promoted to senior engineer in 1952 and subsequently has been manager of radio and carrier development departments and of the former development engineering division. In 1957 he was named product development manager. His previous experience also includes six years of active duty as a naval reserve officer, and three years in development work with the Mackay Radio and Telegraph Co.

Kebby has the B.S. degree in electrical engineering from Stanford University.

(*Continued on page 50A*)

TV camera at distant gate scans person seeking admission to refinery.

How *Foamflex Coaxial Cable*

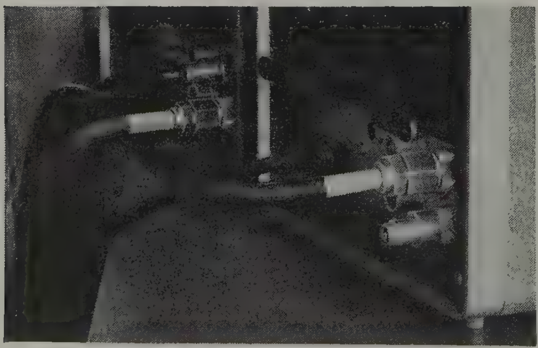
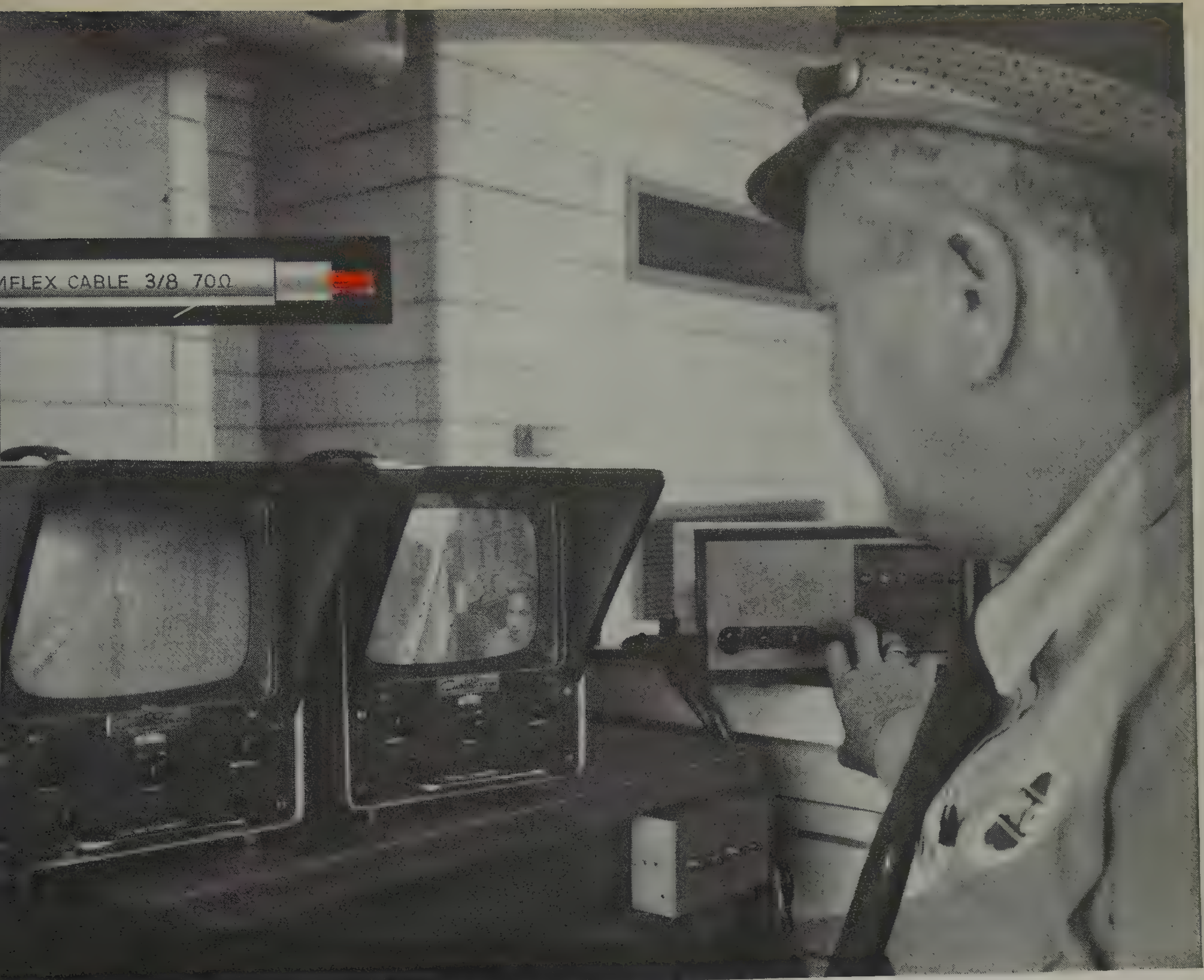


Photo shows Foamflex cable terminations at TV monitor sets in main gatehouse.

Phelps Dodge high frequency cable used in remote control TV monitoring system

Electronic eyes have replaced human eyes at the gates of the Texas City, Texas, refinery of the American Oil Company. The main link of the new closed-circuit television identification and admission system is 11,000 feet of



Guard at main gate checks identity via TV monitor set and intercom set connected with distant gate.

elps guard AMOCO refinery...

" 70 ohm Foamflex coaxial cable with a abirlene (polyethylene) sheath permitting rect burial.

This Foamflex cable is used to join TV meras at remote gatehouses with TV moni- r sets in the control center; the audio circuit r the system is wired with Phelps Dodge xchange Area (direct burial type) telephone ble. From the control center, a guard can entify anyone seeking admission at the re- nery's distant gates. By use of remote switching equipment, tied into the telephone cable, he can also open and close the gates.

Foamflex's low loss, long operating life, and low noise to high signal level ratio are particularly suited to all types of television monitoring operations, including visual control of critical instruments.

Perhaps this Foamflex application suggests a way in which you can use this versatile cable to answer your problems. For further information, write Dept. FC.

PHELPS DODGE COPPER PRODUCTS

CORPORATION

300 Park Avenue, New York 22, N. Y.

IRE People

(*Continued from page 47A*)

He is a member of the American Management Association and an associate member of the American Institute of Electrical Engineers.

❖

Humboldt W. Leverenz (SM'54–F'56) has been announced as the new director of research of the RCA Laboratories. Mr. Leverenz, a pioneer in research on luminescent materials, has been assistant director of research since January, 1958. He has been associated with RCA since 1931, when he joined the corporation in Camden, N. J., as a chemico-physicist. In 1942, he transferred with RCA's research operations to the new research center at Princeton, where he assumed supervision of research on electronically active solid materials. He was named director of the physical and chemical research laboratory in 1954. In his new position, he will be responsible for the general administration of the RCA Labs. research program at the David Sarnoff Research Center in Princeton, and at laboratories at Riverhead and Rocky Point, New York, and Zurich, Switzerland.

In 1940 Mr. Leverenz received the Modern Pioneer Award of the National Association of Manufacturers. He was

(*Continued on page 52A*)

85′ diameter tracking antenna, shown under construction. Reflector face surface is fabricated from aluminum. Pedestal, Polar Cage, Declination Cage and back up structure are of galvanized steel.

IRE People

(*Continued from page 50A*)

awarded the Frank P. Brown Medal of the Franklin Institute in Philadelphia in 1954 for his contributions to the development of the fluorescent lamp. He is the author of numerous scientific and technical articles as well as a basic work, "An Introduction to Luminescence of Solids," published in 1950.

A native of Chicago, Mr. Leverenz graduated from Stanford University and studied physics and chemistry at the University of Munster, in Germany, as an exchange fellow of the Institute of International Education. He is a Fellow of the American Physical Society and a member of the American Chemical Society, the American Association for the Advancement of Science, Sigma Xi, Phi Lambda Upsilon, and Alpha Chi Sigma.

❖

General Electric's Semiconductor Products Department has established a Rectifier Product Section with **C. Graydon Lloyd** (M'45–SM'51) of Skaneateles, N. Y., as general manager.

C. G. Lloyd

A native of Toronto, Canada, Mr. Lloyd graduated from the University of Toronto in 1935 with a B.S. degree in electrical engineering. That same year, he joined the Canadian General Electric Co. and later held marketing and engineering management positions in the C.G.E. Electronic operation.

He was transferred to the company's former Electronics Division in Syracuse in 1952 as assistant to the general manager of the former Commercial Products Department. He was later appointed manager of engineering for the Technical Products Department. He served in that position until 1957 when he was appointed general manager of the Specialty Electronic Components Department in Auburn.

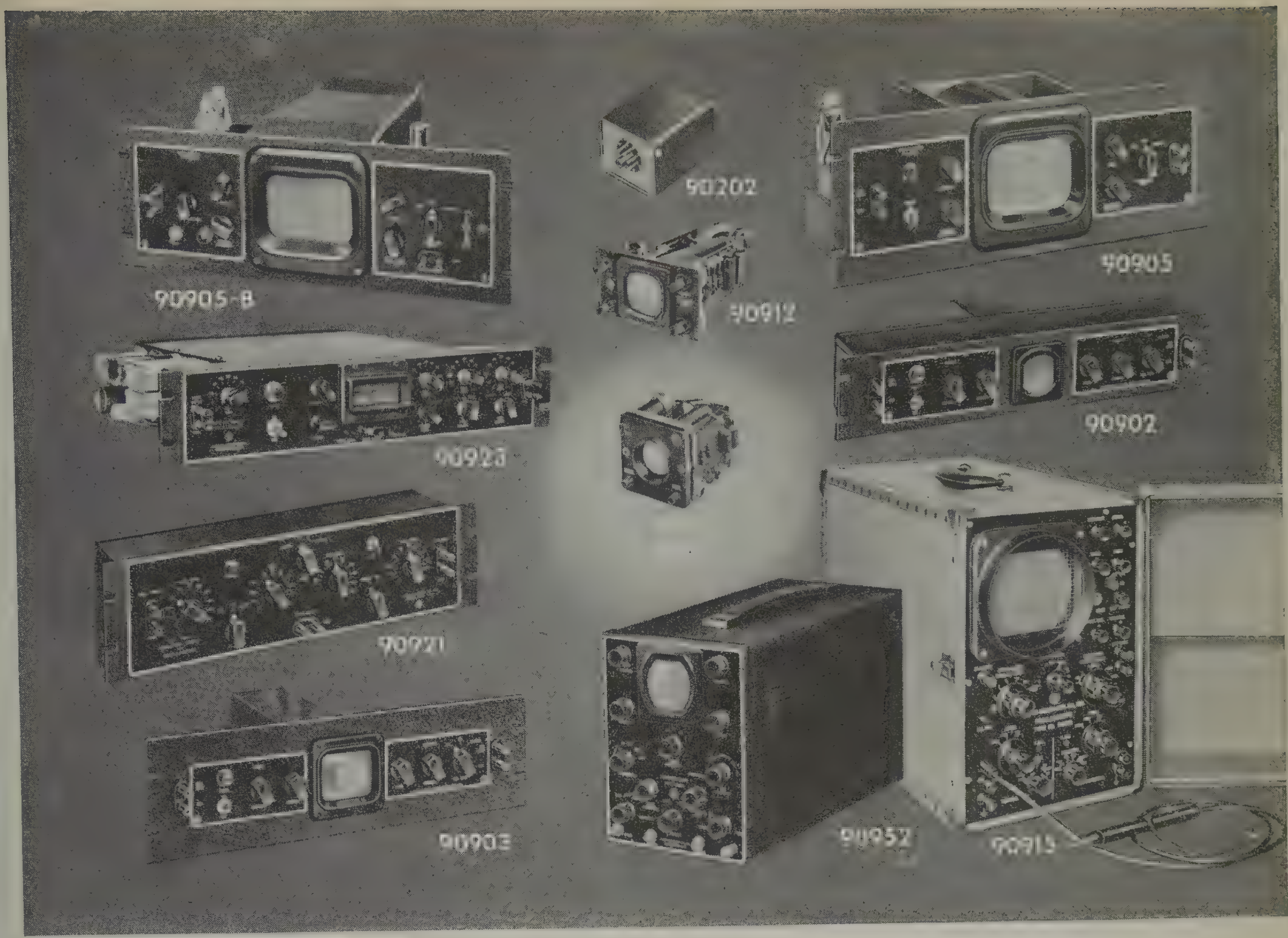
From 1941 to 1945 Mr. Lloyd served in the Royal Canadian Navy, where he held the rank of Lt. Commander in charge of Radio Communications Engineering.

He has served as chairman of the IRE Syracuse Chapter of the Professional Group on Engineering Management. He is a member of the American Institute of Electrical Engineers.

❖

Mr. Maurice Parisier (A'42), president of the electronics and industrial firm of Maurice I. Parisier and Co., has recently been awarded the Cross of "Chevalier de la Légion d'Honneur" in recognition of his services to the development of French radio technique in the United States.

(*Continued on page 54A*)



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The extensive Millen "Designed for Application" line of oscilloscopes and accessories includes five instrumentation oscilloscopes, six rack mounted basic oscilloscopes, an insulated industrial oscilloscope, a miniature synchroscope-oscilloscope, two compact rack mounted complete oscilloscopes, a rack mounted basic oscilloscope for military applications, two amplifier-sweeps, and a plug-in power supply.

INSTRUMENTATION OSCILLOSCOPES

Miniaturized, packaged panel mounting cathode ray oscilloscopes designed for use in instrumentation in place of conventional "pointer type" moving coil meter. Magnitude, phase displacement, wave shape, etc. are readily displayed.

No. 90901 uses type 1CP1 fixed focus one inch cathode ray tube. $2^{11}/_{32}''$ x $2\frac{3}{4}''$ panel. Panel bezel matches in size and type the standard 2" square meters.

No. 90911 uses Type 1EP1 cathode ray tube. Balanced deflection. Blanking input. Sharp focus. Panel matches 2" square meters. Flat face RCA $1\frac{1}{4}''$ diameter tube.

No. 90912 uses type 2BP1 two inch cathode ray tube. 3" x 5" panel. Sharp focus. Good sensitivity. Accelerating voltage 500 to 875 volts. Min. control interaction.

No. 90912-R uses type 3UP1 $2\frac{1}{8}''$ x $1\frac{9}{16}''$ rectangular cathode ray tube.

No. 90913 uses type 3XP1 3" x $1\frac{1}{2}''$ rectangular cathode ray tube. $1\frac{1}{8}''$ x $2\frac{3}{4}''$ useful scan. Vertical sensitivity 33 volts d.c. per inch at 2000 v. accelerating.

BASIC OSCILLOSCOPES

Rack mounted inexpensive *basic* oscilloscopes including cathode ray tube circuit, power supply; intensity, focus, and centering controls, magnetic shielding, safety features, switches, etc. The *basic* oscilloscopes in their packaged form are entirely adequate for many laboratory as well as industrial and communication uses.

No. 90902 uses type 2BP1 two inch cathode ray tube. $3\frac{1}{2}''$ x 19" panel. Power supply — 105–125 volts — 60 cycles. Power consumption — 19 watts.

No. 90902-M — Military version of 90902.

No. 90903 uses type 3KP1 three inch cathode ray tube. $5\frac{1}{4}''$ x 19" panel. Power supply — 105–125 volts — 60 cycles. Power consumption — 19 watts.

No. 90903-R uses type 3XP1 3" x $1\frac{1}{2}''$ rectangular cathode ray tube. $3\frac{1}{2}''$ x 19" panel. Power supply 105–125 volts — 60 cycles.

No. 90905 uses type 5UP1 five inch cathode ray tube. 7" x 19" panel. Power supply 105–125 volts — 60 cycles. Power consumption — 32 watts.

No. 90905-B uses type 5ADP1 five inch flat face precision tolerance cathode ray tube. Power supply — 105–125 volts — 60 cycles. Power consumption — 35 watts.

No. 90905-R uses type B1204 $4\frac{5}{8}''$ x $2\frac{5}{8}''$ rectangular cathode ray tube. $3\frac{1}{2}''$ x 19" panel.

INDUSTRIAL OSCILLOSCOPE

Suitable for use in factory, laboratory, and the field for design, installation, maintenance, and service. Completely insulated front panel and case. Double shielded against magnetic fields. Excellent linearity. Sharp focus over entire 4" x 4" useful scan. The vertical and horizontal amplifiers are stable d.c. amplifiers and are identical, thus permitting accurate phase measurements.

No. 90915 uses type 5AQP — 1, 2, 7, or 11 flat face, precision tolerance cathode ray tube. Frequency response of either amplifier D.C. to 100 K.C. + 0–10%.

MINIATURE SYNCHROSCOPE-OSCILLOSCOPE

Compact "field service" Synchroscope or Oscilloscope. $7\frac{1}{2}''$ x $5\frac{5}{8}''$ x 13". Weighs 17 pounds. Synchronizes to internal or external positive or negative pulses. Band width 10 cycles to 1000 KC. Sweep 6 to 300 microseconds per inch. Performance has not been sacrificed in designing this unit for light weight.

RACK MOUNTED OSCILLOSCOPES

Complete with amplifiers and sweep. Good low frequency response and linearity. For monitoring, production test, or laboratory use. Compact.

No. 90923 uses type 3XP1 3" x $1\frac{1}{2}''$ rectangular cathode ray tube. $3\frac{1}{2}''$ x 19" panel.

No. 90925 uses type B1204 $4\frac{5}{8}''$ x $2\frac{5}{8}''$ rectangular cathode ray tube. $3\frac{1}{2}''$ x 19" panel.

PLUG-IN POWER SUPPLY

Compact high voltage power supply for oscilloscopes, etc. 2" x $2\frac{1}{2}''$ x 5". Input 117 volts 50/60 cycles at 10 watts. Output 750 volts d.c. at 3 ma. and 6.3 volts a.c. at 600 ma. Supplies accelerating and centering potential for oscilloscopes.

No. 90202 Power Supply for Instrumentation Oscilloscopes.

AMPLIFIER/SWEEP UNITS

Horizontal and vertical amplifiers and sawtooth sweep generator for use with *basic* oscilloscopes. Match MILLEN *basic* oscilloscopes in appearance.

No. 90921 — 6SJ7 amplifiers. 6SN7-GT hard tube sweep. $5\frac{1}{4}''$ x 19" rack panel.

No. 90922 — $3\frac{1}{2}''$ x 19" rack panel. Good low frequency linearity.

JAMES MILLEN MFG. CO., INC.

MAIN OFFICE AND FACTORY

MALDEN, MASSACHUSETTS, U.S.A.

IRE People

(Continued from page 52A)

Mr. Parisier holds degrees in mechanical engineering from Kyfhauser Polytechnikum in Germany, in electrical engineering from the Institut d'Electricité de Grenoble, and in electronics engineering from the Ecole Supérieure d'Electricité in Paris, France. He was also given a Certificate from the Law School of the University of Grenoble.

From 1927 to 1941 Mr. Parisier was employed by various industrial firms as design and development engineer and chief engineer for research, construction and installation of communications equipment. During this time he also obtained numerous French, German, Italian, English and United States patents. He was a member of the Commission of Telecommunications of High Voltage Network of the French Ministry and was also a member of numerous important French technical missions from 1938–1941. He was technical consultant to the Delegation in the United States of the "Comité Francaïs de la Libération Nationale" and later continued in this capacity with the French Ministry of Industrial Production.

Mr. Parisier has been an associate member of the French Chamber of Commerce of the United States for the past ten years.

❖

Robert L. Plouffe (S'50–A'51–SM'56), has been named associate laboratory director of the Radio Communications Lab. of the International Telephone and Telegraph research division in Nutley, N. J., and **Harold Havstad** (A'39–VA'39–SM'57) has been appointed an executive engineer.

Mr. Plouffe, a Navy veteran, received a B.S. degree in electrical engineering from Massachusetts Institute of Technology. He also attended the Polytechnic Institute of Brooklyn. He joined the ITT Systems in 1951, and prior to his promotion he was an executive engineer engaged primarily in data processing projects.

A former senior project engineer, Mr. Havstad moves up to the position vacated by Mr. Plouffe. His background includes participation as a radio engineer in a 1928 North Pole expedition and pioneer radio installation work in all parts of the world. An electrical engineering graduate of Oslo University, in his native Norway, he has been with ITT since 1948.

❖

The President of Iowa State College and the Iowa State Board of Regents have announced the appointment of **George R. Town** (A'37–SM'44–F'50) as the dean of the Division of Electrical Engineering and Director of both the Iowa Engineering Experiment Station and the Engineering Extension Service.

Mr. Town was born in Poultney, Vt., on May 26, 1905. He received the degree of E.E. in 1926 and the D.Eng. degree in

(Continued on page 56A)

NEW HORIZONS...

...in Mu Metal, Aluminum, Steel, Stainless Steel, Brass, Copper and Nickel-Silver

There is no substitute for Hudson quality and service in the manufacture of instrument cases ranging from sub-miniature transistor closures to large transformer housings. Here, large scale, continuously modernized facilities are geared with the latest production techniques to meet your most rigorous requirements promptly, efficiently and economically. In addition Hudson offers over 1500 standard items for use "as is" or with modifications. These, along with Hudson's MIL-T line, offer speed of delivery and potential economies too important to overlook on any closure job. Why not investigate today — then do as most companies across the nation have done — *make Hudson your first source of supply for quality closures and covers.* Send drawings or "specs" for quotations on relay closures, transformer housings, transistor and diode closures, instrument cases, sub-assemblies, metal stampings, etc.

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(Continued from page 54A)

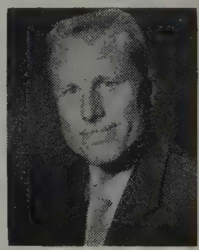
1929 from Rensselaer Polytechnic Institute, where he later served one year as an instructor of mathematics and two years as an instructor in electrical engineering.

He has been affiliated with the research and engineering divisions of several companies including Leeds and Northrup, Arma Engineering Co., and Stromberg-Carlson Co., where he served as an engineer in the research department, engineer-in-charge of the television laboratory, assistant director of research, manager of engineering and research, and in 1945 was elected assistant secretary of the company. He is presently executive director of the Television Allocations Study Organization in Washington, D. C.

Mr. Town has served as member and officer of various panels and committees of the Television System Committee and of the Radio Technical Planning Board. He has been an active IRE member, having served on numerous Institute committees and technical committees and as Director of the Rochester Section. He was elected a Director in 1949 and was made a Fellow in 1950 for his contributions in radio receiver engineering and research.

❖

John C. Riedel (A'52–M'56) has been appointed senior project engineer-electronics for Endevco Corporation, in Pasadena, Calif. He is responsible for engineering development of Endevco amplifiers and other supporting electronic products.



J. C. Riedel

Before coming to Endevco Corporation, Mr. Riedel was test engineer with Consolidated Electrodynamics and supervising engineer of Electronics Group with Northam Electronics.

A native of Pasadena, Riedel, 36, received his B.S. degree in 1947 and his M.S. degree in 1948 from California Institute of Technology.

❖

Dr. Robert D. Teasdale (S'45–A'46–M'49–SM'52) has been named assistant head of the systems analysis department of Hughes Aircraft Company's Ground Systems Group, it was announced today by Dr. Nicholas A. Begovich, director of engineering.

Dr. Teasdale joined Hughes from Melpar, Inc., of Falls Church, Va., where he had been staff assistant to the chief engineer.

From 1952 to 1958, Dr. Teasdale taught at La Salle College in Philadelphia and in 1956 was appointed chairman of the mathematics department of the college's

(Continued on page 58A)

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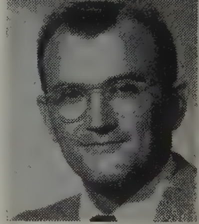
evening division. He previously was associate professor of electrical engineering at Georgia Institute of Technology and a lecturer at Illinois Institute of Technology.

Dr. Teasdale holds a bachelor's degree from Carnegie Institute of Technology, master's and doctor's degrees from Illinois Institute of Technology, all in electrical engineering, and a bachelor of law degree from John Marshall Law School.

He is a member of the American Management Association, Sigma Xi, American Ordnance Association, and the AIEE. He has published a number of engineering papers in the fields of wave propagation, magnetic material, oscillator design and others.

❖

Charles H. Single (A'53) has been appointed computer engineering manager for the Berkeley Division of Beckman Instruments in Richmond, Calif. He moves up from a position as chief project engineer responsible for technical design of the company's EASE computer. Mr. Single's experience in computer design began at Michigan State College where he wrote a master's thesis in electrical engineering on analog computers.



C. H. Single

Before joining the Berkeley Division in 1955 he worked as senior scientist with the Physics Group of the Westinghouse Atomic Power Division. There he was responsible for testing the design of the *Nautilus* submarine with analog computing and simulating techniques. He is a member of the American Association for Computing Machinery.

❖

Dr. Victor Twersky (S'47–A'48–M'55) has been appointed to the staff position of laboratory consultant at the Electronic Defense Laboratory of Sylvania Electric Products Inc.

Dr. Twersky, who is internationally known for his work in the fields of electromagnetic wave scattering and propagation, joined Sylvania in 1953. Before coming to Sylvania, he spent more than three years as a staff member of the Institute of Mathematical Sciences, New York University, and as a consultant in wave propagation at the Nuclear Development Associates in New York. His earlier work included development of electro-acoustic and mechanical guidance devices to aid the blind in foot-travel and obstacle avoidance, and studies on the physical basis of obstacle perception by audition.

(Continued on page 60A)


BUSS
FUSES

(Continued from page 58A)

Dr. Twersky received the B.S. degree from The City College of New York, the M.A. degree from Columbia University, and the Ph.D. degree from New York University.

He is a Fellow of the American Physical Society, a member of the New York Academy of Sciences, President of the Sequoia branch of the Scientific Research Society of America, as well as a member of various other scientific organizations. Also, he was a National Research Council Delegate to the General Assemblies of the International Scientific Radio Union (URSI) in 1954 and 1957, and is presently an official member of Commission VI of URSI.

❖

The appointment of **Dr. Sherrerd B. Welles** (SM'56) as senior engineering specialist for Sylvania Electronic Systems, a major division of Sylvania Electric Products Inc., has been announced by Dr. Edwin G. Schneider, chief engineer of Sylvania Electronic Systems.

Dr. Welles will be responsible for improving and maintaining the interchange of technical information, both within the division and between Sylvania Electronic Systems and other divisions of the corporation. He will make his office at the division headquarters in New York City, N. Y.

Dr. Welles joined Sylvania's Missile Systems Laboratory in 1954 as manager of the electronics department and, in 1955, he became manager of the Missile Systems Laboratory. He left Sylvania early last year to join Lockheed Aircraft Corp.

Prior to joining Sylvania, he was with the Massachusetts Institute of Technology Radiation Laboratory, the Air Force Cambridge Research Center, and also was a member of the Air Defense Systems Engineering Group.

Dr. Welles is a graduate of Yale University with a B.S. degree in mathematics and a Ph.D. in physics. He is a member of Phi Beta Kappa and Sigma Xi, as well as the American Physical Society and Research Society of America.

❖

A new position carrying the title of Fellow, Technical Staff, has been established by RCA Laboratories to recognize the continued outstanding individual achievement in the field of research, it was announced by **Dr. Irving Wolff** (A'27–VA'39–F'42), Vice-President of Research. The announcement included the designation of eight RCA Labs. scientists to the new position of Fellow, Technical Staff, and the naming of three others as associate laboratory directors. In announcing the creation of the new position, Dr. Wolff explained that it is intended to give recognition to outstanding individual contributions, just as recognition is extended by

(Continued on page 62A)

IRE People

(*Continued from page 60A*)

title and position for those made through group administration. He compared the new designation with the similar use of the title by virtually all technical societies.

"We consider this appointment a badge of high technical achievement," he said. "We hope that the title of Fellow will confer the same recognition as that of Associate Laboratory Director. The difference will be mainly of personal volition: the Fellow desiring to make continued personal technical contributions, and the Associate Laboratory Director and Laboratory Director electing to contribute through leadership in group administration."

Among the eight RCA Labs. scientists designated as Fellows are **Alda V. Bedford** (A'31–SM'46–F'50), **Clarence W. Hansell** (A'26–M'29–SM'43–F'53), **Ray D. Kell** (A'35–F'47), **Nils E. Lindenblad** (M'34–SM'43–F'48), **Dwight O. North** (A'35–M'38–SM'43–F'43), **Edward G. Ramberg** (M'53–SM'53–F'55), and **Albert Rose** (A'36–M'40–SM'43–F'48). The three new associate laboratory directors are **Harwick Johnson** (SM'45), **Leon S. Nergaard** (A'29–M'38–SM'43–F'52), and **Jan A. Rajchman** (SM'46–F'53).

Stating that additional Fellows will be appointed from time to time, Dr. Wolff said that their selection, like those of Associate Laboratory Directors, will be based on a set of criteria including research achievement, creativity, technical leadership, and recognition both in and outside RCA for vision, judgment, and knowledge.

Section Meetings

Akron

"Light and Architecture at the Brussels World Fair," K. Staley, G.E. Co.; 1/21/59.

Alamogordo-Holloman

Field Trip to High Speed Test Track, Lt. Col. Vlcek, Maj. Nolte, and Staff, Air Force Missile Development Center; 1/19/59.

Albuquerque-Los Alamos

Social Evening; 12/13/58

"Telemetry Receiver Threshold Improvement," K. A. Gilchrist, Jet Propulsion Lab.; 1/20/59.

Anchorage

"Railroad Communications in Alaska," F. W. Shellhorn, Alaska Railroad; 1/5/59.

Business Meeting; 2/2/59.

"The Information Machine," J. G. Tryon, Univ. of Alaska: 2/9/59.

Baltimore

"Compatible Color Television," J. Wentworth. RCA; 1/14/59.

BAY OF QUINTE

"IGY and a Visit to USSR," G. A. Harrower, Queen's Univ.; 1/20/59.

"Flight Test System for DC-8," R. L. Bentley, Consolidated Electro-Dynamics Corp.; 1/26/59.

BUFFALO-NIAGARA

"Theory, Applications of Parametric Amplifiers and Masers," L. Eastman, Cornell Univ.; 2/11/59.

CEDAR RAPIDS

Annual Dinner Meeting and Installation of Officers; 1/17/59.

CHINA LAKE

Annual Business Meeting; 12/4/58.

CLEVELAND

"Some Properties and Uses of Artificial Dielectric Media," R. E. Collin, Case Institute of Technology; 1/15/59.

COLUMBUS

"The Role of BOMARC in Air Defense," R. D VanNest, Boeing Airplane Co.; 2/10/59.

DALLAS

Presentation of Fellow Awards to R. W. Olson and J. R. Macdonald, C. E. Harp, Univ. of Okla.; 1/30/59.

Discussion and Demonstration of Stereodisk Recordings, T. W. Swafford and A. C. Morris, Hi-Fi-Tron Associates; 2/2/59.

DAYTON

"Electronics in Solids, Space and Sound," C. N. Hoyler, RCA; 12/11/58.

DENVER

"Evolution of Instrumentation," W. Griffith, Martin Co.; "System Approach to Data Handling in Ballistic Missile Development Program," D. A. Rodgers, Ramo-Wooldridge; "Installation and Operation of a Large Central Data System," R. Young, Martin Co.; 1/22/59.

EGYPT

"The Application of an Electrodeless Discharge in the Design of Ion Sources," M. E. Abdel Aziz, Ein-Shams Univ.; 2/4/59.

ELMIRA-CORNING

"Building the Mackinac Straits Bridge," Movie and Talk, E. F. Green, American Bridge Co.; Joint meeting with NYSPE, ASTE and SAE; 12/5/58.

"Project Matterhorn," G. Lewin, Princeton Univ.; 1/19/59.

EMPORIUM

"Closed Circuit TV," Demonstration and Lecture, D. Swaine, Sylvania Electric Prod. Inc.; 1/20/59.

"Video Tapes and How They Work," C. Luscombe, Ampex Corp.; 2/17/59.

EVANSVILLE-OWENSBORO

"Single Side-Band," W. B. Bruene, Collins Radio Co.; 1/14/59.

FLORIDA WEST COAST

"High Reliability at Low Cost," B. L. Weller, Vitramon, Inc.; 1/21/59.

FORT HUACHUCA

"Ever Increasing Role of Electronics in Drone Development," W. E. Peterson, Radioplane; "Drone Tracking System with Lightweight Airborne Package," E. Walcek, Radioplane; 1/19/59.

"Parametric Amplifiers," A. L. Aden, Motorola; 2/16/59.

(Continued on page 64A)

Section Meetings

(*Continued from page 63A*)

HAMILTON

"Scatter Propagation," Mr. Lipinski, Canadian Westinghouse Ltd.; 1/9/59.

"Klystrons—A Survey of Modern Trends," J. Matties, Varian Associates of Canada, Ltd.; 2/9/59.

HOUSTON

"Ultraviolet Flying—Spot Television Microscope," P. O'B. Montgomery and W. A. Bonner, Southwestern Medical School; 1/20/59.

"Seismic Data Reduction Systems," A. Smith, Humble Oil & Refining Co.; C. Frobese, Electro-Tech Labs.; L. McManis, SIE; F. Rotramel, Elecdrodynamic Instrument Co.; B. Tickell, Clevite Corp.; H. J. Jones, Texas Instruments; Joint with PGI; 2/17/59.

HUNTSVILLE

"The Role of Radio Telemetry in Space Exploration," H. Scharla-Nielsen, Radiation Inc.; Election of PGTRC Officers; 1/30/59.

INDIANAPOLIS

"Technical Analysis of Home Electrical Equipment," W. Ayres, Howard W. Sams & Co.; 2/5/59.

ISRAEL

"Ferro Resonance," A. Wulkan, Israel Ministry of Defense; 11/30/58.

ITHACA

"Recent Advances in Medical Electronics," V. Zworykin, Medical Electronics Center of Rockefeller Institute; 1/19/59.

"Magnetohydrodynamics and Its Application to Space Propulsion," A. Kantrowitz, Avco Mfg. Corp.; 2/9/59..

KANSAS CITY

"The Pade Approximation for the Delay Operator," C. Halijak, Kansas State College; Business Meeting; 2/10/59.

"Jet Age Navigation Systems," H. K. Morgan, Bendix Aviation Corp.; 1/13/59.

LITTLE ROCK

"Resistors," T. Pfister, Allen Bradley; 1/7/59.

LONG ISLAND

Introduction, A. Kantowitz, AVCO Research Lab.; 1/8/59.

"Fusion Physics," J. Johnson, Princeton Univ.; 1/15/59.

"Thermonuclear Devices," S. Koslov, Stevens Institute; 1/22/59.

"Oscillations in Plasmas," J. Berkowitz, New York Univ.; 1/29/59.

LOS ANGELES

"Submarines—Past, Present and Anticipated Future," R. E. Styles, Submarine Flotilla I; "Trans-Polar Crossing of Submarine *Nautilus*," S. Jenks, *Nautilus* Submarine; a Joint Meeting with Santa Ana Subsection; 1/8/59.

"Pacific Missile Range," S. R. Radom, U. S. Naval Air Missile Test Center; "Whither Transistor Electronics," W. S. Shockley, Shockley Semiconductor Lab.; Joint Meeting with Santa Barbara Subsection; 12/5/58.

"Medical Problems Related to the Nuclear Power Field," S. Warren, School of Medicine, UCLA; "Nuclear Power," C. Starr, Atomics International; Joint meeting with AIEE; 11/4/58.

(*Continued on page 66A*)

ALL-PURPOSE DIGITAL VOLT-OHM METER

Examine these outstanding features.

RAPID, ERROR-FREE READINGS BECAUSE

Type of measurement indicated.

Polarity automatically displayed.

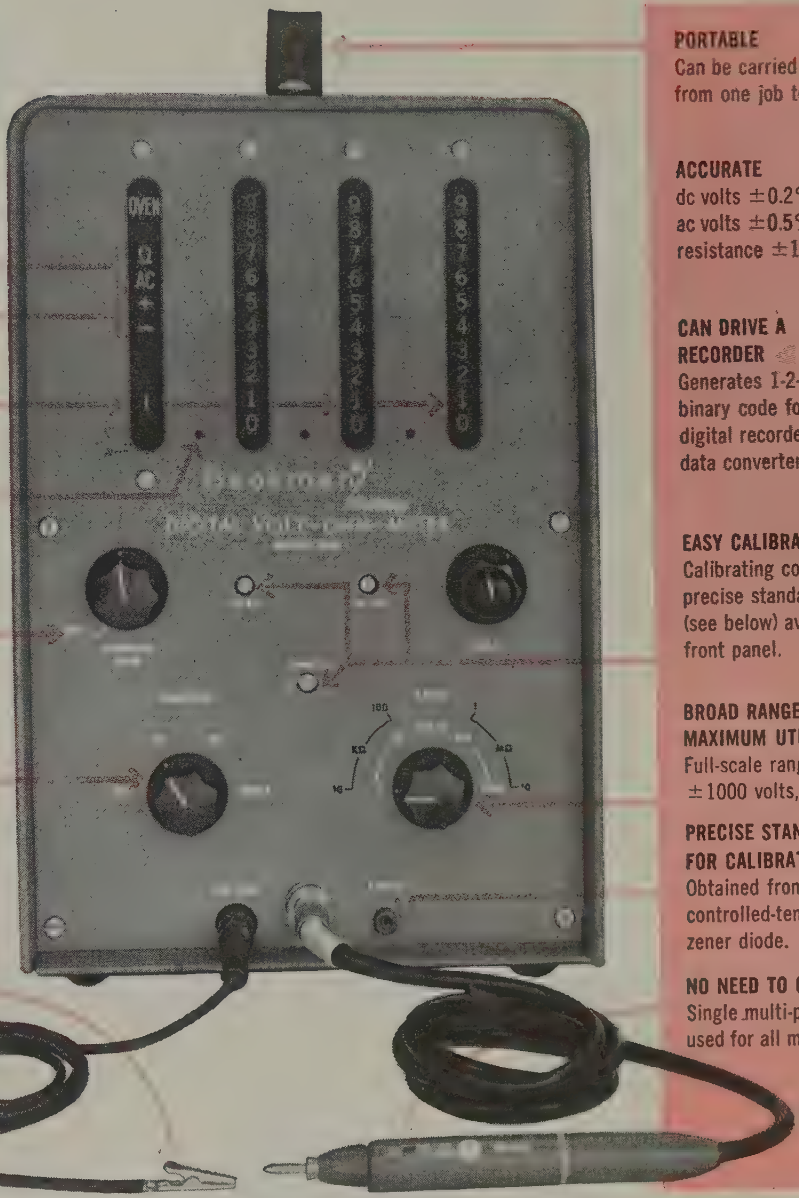
Digital display. No multi-scale confusion, interpolation or parallax error.

Decimal point automatically positioned.

PROVISION FOR REMOTE CONTROL
Measurements can be triggered by external command signal.

THREE TYPES OF MEASUREMENT
dc volts, ac volts and resistance.

SIGNAL GROUND ISOLATED FROM CHASSIS
You can measure voltage between two points when neither is at ground.



160

PORTABLE
Can be carried easily from one job to another.

ACCURATE
dc volts ±0.2%
ac volts ±0.5%
resistance ±1.0%

CAN DRIVE A RECORDER
Generates 1-2-2-4 binary code for digital recorder, data converter, etc.

EASY CALIBRATION
Calibrating controls and precise standard voltage (see below) available on front panel.

BROAD RANGE INSURES MAXIMUM UTILITY.
Full-scale ranges of ±1 to ±1000 volts, 10k to 10M ohms.

PRECISE STANDARD VOLTAGE FOR CALIBRATION
Obtained from an internal controlled-temperature zener diode.

NO NEED TO CHANGE PROBES
Single multi-purpose probe used for all measurements.

As the picture reveals, BECKMAN/Berkeley's Model 5350 is the most useful, most versatile digital instrument of its kind. It offers operating flexibility and features not found in digital voltmeters costing three times as much. The Model 5350 makes it feasible to replace multi-purpose analog equipment with a more accurate, rapid and foolproof means of making the vast majority of everyday voltage and resistance readings.

Three digits present all readings within the nominal full scale range (000 to 999), a fourth digit permits off-scale readings up to 150% of full scale. All electronic construction eliminates troublesome stepping switches and permits an instantaneous display of readings at rates up to 10 per second.

Priced at only $845.00.

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Berkeley Division
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a division of Beckman Instruments, Inc.

Section Meetings

(Continued from page 64A)

LUBBOCK

"Video Recordings," J. Burge, KDUB-TV; Ampex VR 1000 Tape Recorder, a demonstration; 1/6/59.

MIAMI

"Telemetering," 1958 National Symposium and Meeting; 9/22/58.

"New Horizons in Electronics," C. N. Hoyler, RCA; 10/21/58.

Talks, Mr. Donald Fink and Mr. Cole; 12/1/58.

"Why Satellites? What Are They Intended to Do?" A. P. Smith, Jr., WKAT; 1/9/59.

Tour, Florida Power & Light Co. plant facilities; 2/13/59.

"Semi-Conductor Development, Production & Application," G. Frieman, Raytheon; B. McCarthy, Raytheon; 2/19/59.

MILWAUKEE

"Satellite Instrumentation for Meteorological Observations," V. E. Suomi, Univ. of Wisconsin; 1/20/59.

MONTREAL

"Operational Research on Radar Performance," G. R. Lindsay, Air Defense Command Headquarters, RCAF; "Mid-Canada Line Doppler Detection System," M. F. Lemke, RCAF; 11/19/58.

"The Amazing Maser," Dr. Marr, McGill Univ.; 12/9/58.

"Basic Digital Computers," R. Harvey, Canadair, Ltd.; 12/17/58.

"Tropospheric Scatter Radio Systems," D. J. McDonald, Bell Tel. Co. of Canada; 1/20/59.

NEWFOUNDLAND

Christmas Social; 12/13/58.

"Ionospheric Measurements Systems," C. F. Post, U. S. Army; 1/21/59

NEW ORLEANS

"The Position of The Institute of Radio Engineers in Respect to Other Technical Societies," Dr. Ernst Weber, IRE President; 1/23/59.

NORTH CAROLINA

"Graphic Arts in Electronics," H. L. Shortt, Technigraph Printed Electronics, Inc.; 1/16/59.

NORTHERN NEW JERSEY

"Creativity and the Engineer," A. N. Goldsmith; 11/12/58.

"Electronic Counter-Measures," L. A. deRosa, ITT Labs.; 1/14/59.

NORTHWEST FLORIDA

"Automatic Ground Controlled Approach System," J. Coyle, Gilfillan Bros., Inc.; 1/20/59.

OTTAWA

Armed Services Night: "Naval Communications," F. T. Gillespie, R. J. Legeer, RCN; "History of Sonar up to 1950," M. T. Gardner, RCN; "Underwater Communications," J. W. Elson, RCN; "Some Aspects of Naval Shore Communications Systems," H. Palmer, RCN; 12/4/58.

"The Solid State Maser," R. A. Armstrong, National Research Council; 1/15/59.

OKLAHOMA CITY

"New Developments, New Techniques of Bell Telephone System," K. Peterson, Bell Tel. System; 1/13/59.

(Continued on page 68A)

When d-c generator regulation troubles pile up . . .

Specify Hydro-Aire's new transistorized voltage regulator

Hydro-Aire's new, completely transistorized regulator for d-c generators weighs less than half as much as the conventional carbon-pile regulator that it replaces. Response time is five times better, and operational life is extended to 10,000 hours. The new Hydro-Aire units are physically and functionally interchangeable with MIL-standard carbon-pile regulators – plug directly into existing receptacles.

Other important features of the new Hydro-Aire unit: Has no moving parts, requires no shock mounting, dissipates virtually no heat, requires no forced cooling, adjusts easily to different voltages.

Model 50-029 (shown above) has these characteristics:

Voltage: may be set to any value between 26 and 30 volts, in increments of 0.1 volt.
Temperature limits: –55°C to +71°C
Rated life: 5000 hours without maintenance
Dimensions: 3¼ x 4⅛ x 2⅞ inches
Weight: 1¼ pounds

Additional ratings are available and our laboratories welcome the opportunity to design new devices to your specific requirements.

Write or wire for complete information on Hydro-Aire's generator regulators, and other solid-state devices.

HYDRO-AIRE

Every Basic Airborne System

Section Meetings

(*Continued from page 66A*)

Philadelphia

"Impact of Computers on Modern Society," M. Rubinoff, Univ. of Penna.; 1/21/59.

1959 Fellows' Night, Dr. Ernst Weber, IRE President; 2/7/59.

Phoenix

"Nuclear Reactor at the University of Arizona," M. Wittmeyer, Univ. of Arizona; 1/9/59.

Pittsburgh

"Characteristics and Applications of Some New Semiconductor Devices," A. P. Kruper, Westinghouse Electric Corp.; "A Remotely Controlled Electroluminescent Totalizing Display," R. C. Lyman, Westinghouse Electric Corp.; 10/20/58. Conducted Tour, WTAE studios and facilities; 11/20/58.

"Propu sion Requirements of Space Vehicles," S. Way, Westinghouse Electric Corp.; 2/9/59.

Portland

"The Impact of Solid State Devices on Communication," R. K. Honaman, Bell Tel. Labs.; 1/22/59.

"What is Electronics Engineering," a Panel Discussion, N. Stunkard, Osborne Electronics, R. Pooley, Tekronix, J. Dannenmann, Pac. Tel. & Tel.; Moderator; Cliff Moulton, Tektronix; Trade Show held Jointly with W. C. E. M. A. "Research at O.S.C.," Prof. Stone, Oregon State College; 2/7/59.

Princeton

"The Present Status of World Efforts to Attain a Controlled Thermo Nuclear Reactor," M. B. Gottlieb, C. Stellerator Associates; 1/8/59.

"The Perceptron—A Probabilistic Model of the Brain," F. Rosenblatt, Cornell Aeronautical Lab.; 2/12/59.

Quebec

"Measurements of Solar Corpuscles and Their Relation to Ionosphere Storms," T. R. Hartz, Radio Physics Lab., D.R.B.; 1/19/59.

Rochester

"Patents and Their Relation to the Electronic Industry," R. W. Hampton, Eastman Kodak Co.; 1/22/59.

St. Louis

"Development of Color Television," C. J. Hirsch, Hazeltine Res. Corp.; 12/16/58.

"Some New Horizons in Communications," G. H. Scheer, C. & N. Labs, WADC; 1/20/59.

San Antonio

"Special Semiconductor Devices," R. Stewart; 12/10/58.

"Noise in Semiconductor Devices," R. Petritz; 12/17/58.

"Semiconductor Surface Problems," G. Peattie; 1/7/59.

Social Meeting, Joint with AIEE, ASME, AFCEA and IRE; 1/15/59.

"Acoustic Definition, Its Meaning, Measurement and Anticipated Applications," J. L. Collins, Boner and Lane, and Univ. of Texas; 1/22/59.

San Diego

"Factors Influencing the Use of Single Side Band," E. W. Pappenfus, Collins Radio Co.; 2/3/59.

Schenectady

"Ultrafine Particles and Permanent Magnets," T. O. Payne, Gen. Electric Co.; 2/10/59.

(*Continued on page 70A*)

PHILCO IS **PEOPLE**

From advanced research and development to mass production, installation and servicing of countless electronic products and systems, Philco is people. Here is a closely integrated organization of scientists, engineers, installation and service specialists, ready to meet any challenge for creation of military, industrial and consumer electronics systems.

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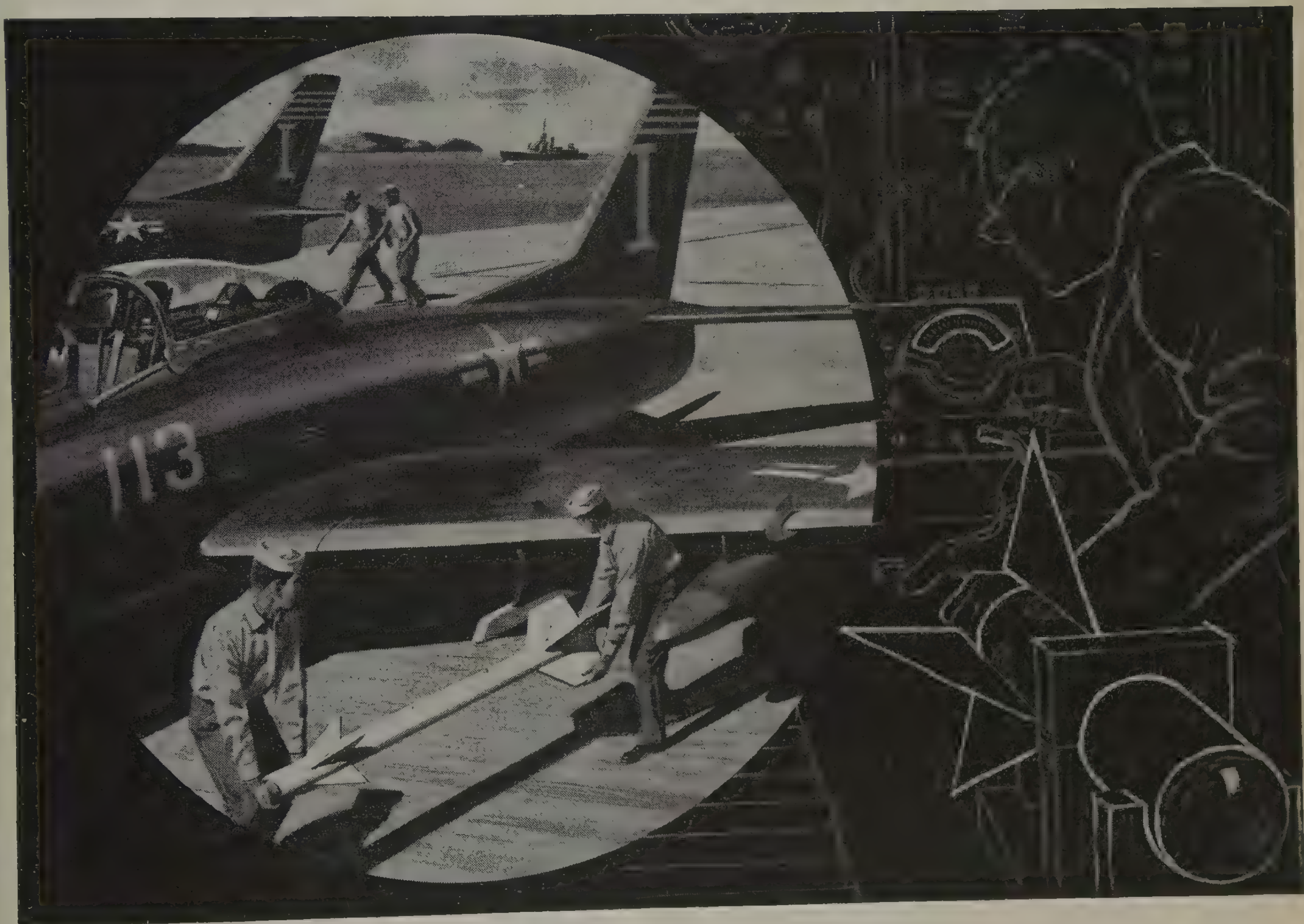

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Section Meetings

(Continued from page 68A)

SOUTH BEND-MISHAWAKA

Tour of the Drewry Plant, Drewry's Ltd.; 12/18/58.

SOUTH CAROLINA

Business Meeting and Nomination of Officers; 12/29/58.

Business and Executive Committee Meetings; 1/23/59.

SYRACUSE

"The Space Age," R. W. Porter, G. E. Co.; a Joint Meeting with PGMIL and the Technology Club of Syracuse; 1/5/59.

VIRGINIA

"Radio Astronomy," N. G. Român, Naval Research Lab.; 1/30/59.

WASHINGTON

General Session, Speakers: C. Hall, Nat'l. Science Foundation, W. J. Boyle, G.E. Co., P. Klass, Aviation Week; a Joint Meeting with PGEWS & PGEM; 2/2/59.

WESTERN MASSACHUSETTS

"The Status of Elementary Particle Research," D. Park, Williams College; 2/6/59.

WICHITA

"Growth of Broadcasting Since the 20's," K. Pyle, KSIR; 1/23/59.

"Single-Side Band and Filter Alignment," L. Dale, Southwestern Bell Tel. Co.; 2/17/59.

WINNIPEG

"Project Janet—Communication System from Meteor Trails," P. Forsythe, Univ. of Saskatchewan; 1/22/59.

SUBSECTIONS

BUENAVENTURA

"Environmental Testing of Explorers," W. S. Shipley, Jet Propulsion Lab; 1/14/59.

EASTERN NORTH CAROLINA

"Global and Satellite Communications," S. G. Lutz, Hughes Aircraft Co.; 1/9/59.

LANCASTER

"The Application of Inertial Guidance Systems for Long Range Missiles," C. J. Mundo, Jr., Am. Bosch Arma Corp.; 1/15/59.

LAS CRUCES

"The Need for Professional Men in the United States," N. Golovin, Office of Tech. Services; a Joint Meeting with American Rocket Soc., Instrument Society of Am., Soc. of Photographic Instrumentation Engineers, and Astronomical Soc. of Las Cruces; 1/23/59.

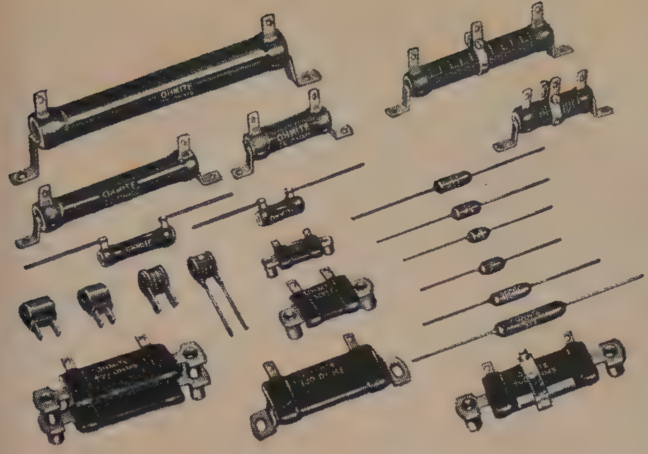
LEHIGH VALLEY

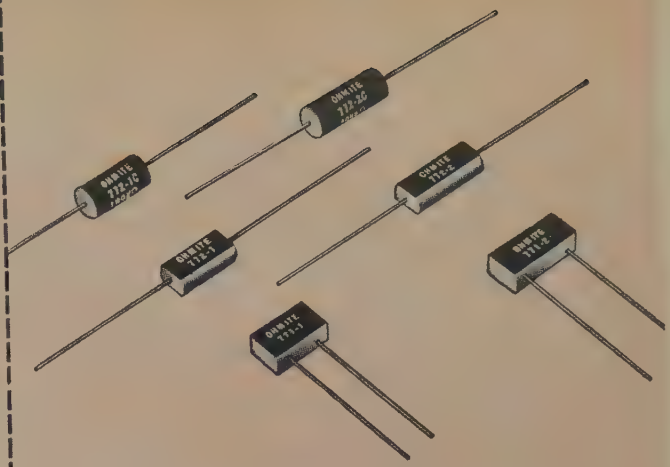
"My Pocket is Calling," K. Ozone, Stromberg Carlson Co., and C. R. Bischoff, Bell Tel. Co. of Pa.; 1/9/59.

MEMPHIS

"Navigational Measurements Methods for Missile Guidance Systems," a Survey, S. L. Johnston, Redstone Arsenal; Election of Officers; 1/30/59.

(Continued on page 72A)

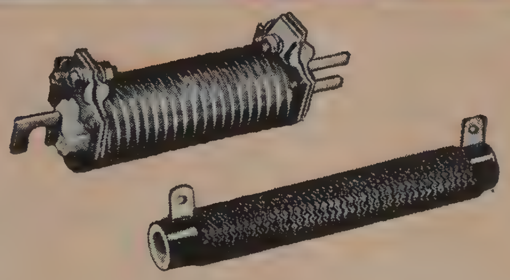


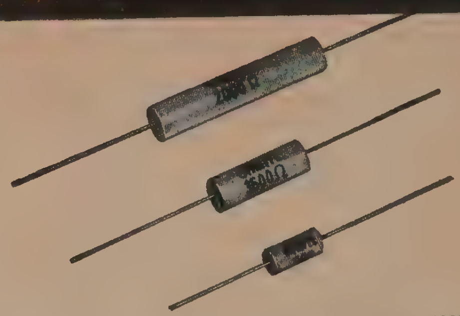


VITREOUS-ENAMELED RESISTORS Tremendous variety of types and sizes. Fixed, adjustable, tapped, noninductive, thin, and precision resistors available in a wide range of wattages and resistances. Also available to meet MIL-R-26C requirements.

METAL FILM RESISTORS Riteohm® metal film precision resistors feature full ¼-watt rating at 150°C ambient; excellent high frequency characteristics; low temperature coefficient of resistance. Long-term load and shelf stability.

OHMITE® INDUSTRY'S MOST COMPLETE LINE OF QUALITY WIRE-WOUND RESISTORS





POWER RESISTORS Power-type resistors for high-current, low-resistance applications. Vitreous-enameled, edge-wound, corrugated ribbon Corrib® units and open-type, edge-wound ribbon or round-wire Powr-Rib® units handle a wide range of power resistor needs. Available in fixed or adjustable "DIVIDOHM®" types.

MOLDED PRECISION WIRE-WOUND POWER RESISTORS Insulated units with Silicone-Type molded covering. Available in 3-, 5-, and 10-watt sizes. Tolerances: 0.1%, 0.25%, 0.5%, 1%, and 3%. Maximum resistance: 3-watt, 10,000 ohms; 5-watt, 25,000 ohms; 10-watt, 50,000 ohms.

OHMITE HAS EXACTLY THE RESISTOR YOU NEED
Ohmite offers the most complete line of high quality resistors on the market . . . fixed, adjustable, tapped, noninductive, and precision resistors in many sizes and types of terminals . . . in a wide range of wattages and resistances. All-welded construction. Ohmite application engineers will be pleased to help you in selecting the resistors for your job.

Write on company letterhead for Catalog 58.

NONINDUCTIVE
Tubular vitreous-enameled resistors with special winding. Dummy antennas consist of assemblies of several resistors. Watts, 5 to 1000; ohms, 1 to 5000.

SPECIAL VARIETIES
Ohmite can provide toroids, flat strips, plaques, special-sized tubes, or tubes with mixed terminals, etc. Watt ratings and resistances available as required.

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RHEOSTATS RESISTORS RELAYS TAP SWITCHES TANTALUM CAPACITORS
R. F. CHOKES VARIABLE TRANSFORMERS DIODES

Section Meetings

(Continued from page 70A)

MERRIMACK VALLEY

"Flood Control in the Merrimack River Basin," A. K. Sibley, U. S. Army; a Joint Meeting with AIEE; 10/20/58.

"Applications of the Electric Arc Plasma Jet," M. E. Malin and R. R. John, AVCO-RAD; Tour of Plasma Jet Facility & Ballistics Range Labs.; 1/19/59.

NORTHERN VERMONT

"Space Technology," J. Bohnslaw, Gen. Electric; a Joint Meeting with AIEE; 1/19/59.

SANTA ANA

"Submarines—Past, Present and Anticipated Future," R. E. Styles, Submarine Flotilla I; "Trans-Polar Crossing of Submarine *Nautilus*," S. Jenks, *Nautilus* Submarine; a Joint Meeting with Los Angeles Section; 1/8/59.

SANTA BARBARA

"Deep Ocean Research," H. Wilcox, Naval Ordnance Test Station; 1/13/59.

Professional Group Meetings

ANTENNAS AND PROPAGATION

Boston—January 7

"Wide-Angle Electronic Scanning," G. Ploussios, Chu Associates.

Boston—February 4

"Microwaves and their Optical Analogues," G. Hull, Sylvania Elec. Corp.

Los Angeles—January 8

"Four Dimensional Antenna Systems," H. H. Shanks, Hughes Aircraft Co.

ANTENNAS AND PROPAGATION MICROWAVE THEORY AND TECHNIQUES

Philadelphia—Feb. 4

"Recent Advances in Microwave Ferrites," E. Wantuch, Airtron, Inc.

AUDIO

Boston—January 14

"Hi-Fidelity Stereophonic Broadcasting," W. S. Halstead, Multiplex Services Corp., R. L. Kaye, WCRB.

Dayton—October 2

"Special Problems Encountered in Doppler-Shift Tracking Earth Satellites," L. W. Root, Geo. Behm and Son's Co.

Milwaukee—January 13

"Four Track Stereo Tape Recorders," D. R. Andrews, RCA.

(Continued on page 76A)

BROADEN DESIGN HORIZONS

with new pnp drift transistors

TYPICAL APPLICATIONS

- TV CIRCUITS
- FM RADIOS
- SHORT WAVE RADIOS
- HIGH FREQUENCY OSCILLATORS
- VERY HIGH SPEED SWITCHING DEVICES

SEVEN NEW DRIFT TRANSISTORS FOR HIGH SPEED SWITCHING AND HIGH FREQUENCY AMPLIFIER APPLICATIONS

General Transistor's new 2N602, 2N603, 2N604 provide the design engineer with guaranteed switching parameters such as gain-bandwidth and DC current gain, while the 2N605, 2N606, 2N607 and 2N608 provide guaranteed power gains at high frequencies.

In addition to the great speed advantages offered by the drift transistor at no sacrifice of gain, such additional features as higher voltages and lower capacity are available. Thus one can now drive higher impedance loads with no sacrifice of speed or pulse power.

The complete control of G. T.'s Drift Transistor assures longer life and maximum performance while possessing complete reliability.

Other features include: high input-circuit efficiency, excellent high-frequency operating stability, good signal-to-noise ratio, good automatic-gain-control capabilities and the rugged mechanical construction of a positive hermetically sealed TO-9 case.

WRITE TODAY FOR BULLETIN G-180 INCLUDING COMPLETE MECHANICAL AND ELECTRICAL SPECIFICATIONS, DIMENSIONAL DRAWINGS, GRAPHS AND ENGINEERING DATA.

ALL TRANSISTORS CAN NOW BE SUPPLIED IN FULL COMPLIANCE WITH MIL-T-19500A.

	DC Current Gain h_{FE}		Gain X Bandwidth		Power Gain K_p		
2N602 2N603 2N604	$V_{CE} = 1$ v $I_B = 0.5$ ma	25-100*	$V_{CB} = 5$ v $I_C = 5$ ma	10-30 mc 30-50 50-70	2N605 2N606 2N607 2N608	$V_{CE} = 7.5$ v $I_C = 1$ ma $f = 2$ mc	20-25 db 25-30 db 30-35 db 35-40 db

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Professional Group Meetings

(Continued from page 72A)

Philadelphia—January 14

"Stereo Transmission by Multiplexed Compatible FM," M. G. Crosby, Crosby Labs., Inc.

San Francisco—January 20

"Demonstration of Wind Quintet Music," M. Ward Widener, Ampex Audio, Inc.

Syracuse—December 2

"Stereophonic Disc Recording," W. S. Bachman, Columbia Records, Inc.

Automatic Control

Long Island—January 20

"Hot Gas Servos and Their Application to Missiles," C. Myer, Sperry Gyroscope Co.

Automatic Control/Production Techniques

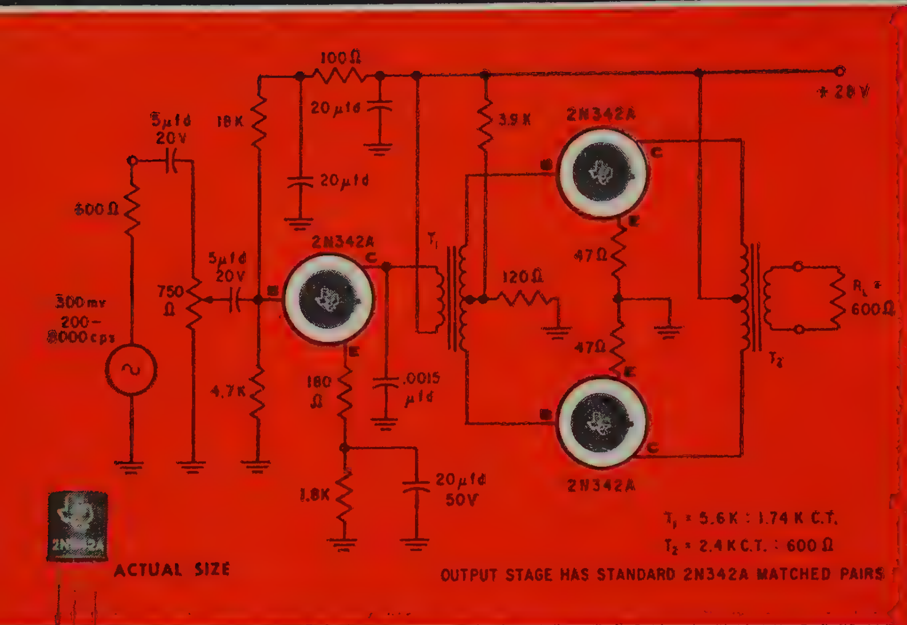
Boston—December 15

Panel Discussion—Some Current Views on Technology in Soviet Russia.

(Continued on page 82A)

TI Silicon Transistor Application Not

TRANSISTORIZED INTERCOM EXCEEDS MIL-E-5272B SPEC



CIRCUIT SPECIFICATIO

150-mw output from -55°C 100°C at less than 10% ha monic distortion over frequen range

Frequency response @ 25 stable within ±2 db of 100 cps 100 mw reference lev from 200 to 8000 cps

Frequency response @ -55 and 100°C within ±3 db 25°C frequency response

Less than 3-db gain variatio @ -55°C and 100°C compare to 25°C measurement

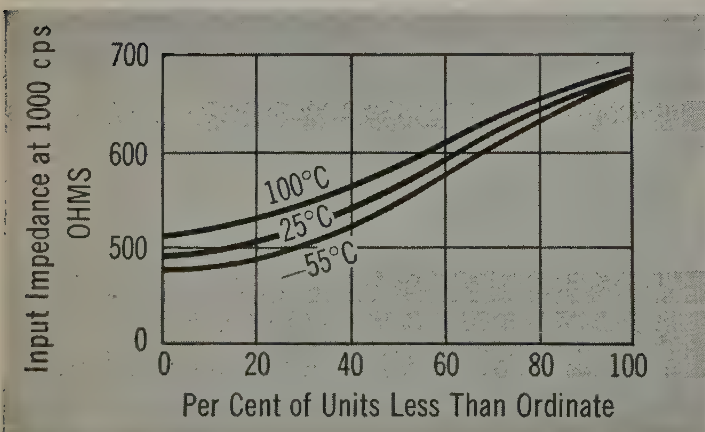
LOUD AND CLEAR AT 100°C!

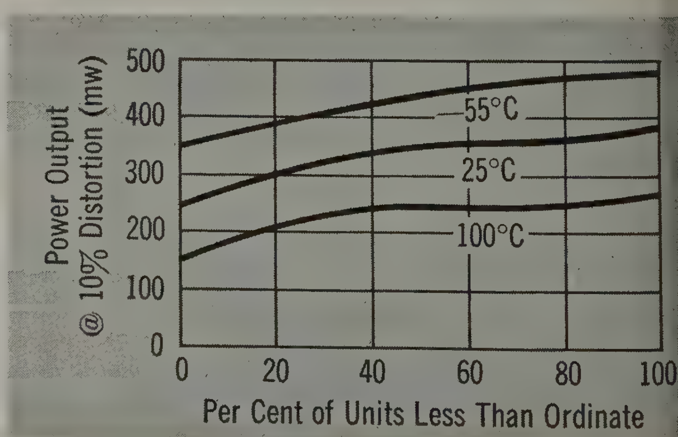
...with TI 2N342A silicon transistors from stoc

You can satisfy the 71°C equipment requirements of MIL-E-5272B at 100°C with the interco amplifier circuit shown above — using TI 2N342A silicon transistors with . . . guaranteed 3-to linear beta characteristics . . . 85-v collector-to-emitter breakdown, giving a wide safety ran with 28-v aircraft supplies . . . plus dissipation capability of 1 watt at 25°C and 200 mw at 125°

The newest addition to the *use-proved* TI 2N339 series introduced in 1957, this medium-pow unit carries the full-year TI guarantee and is immediately available *off-the-shelf* from all distributors in 1-249 quantities. For production quantities, contact your nearest TI sales offi

TYPICAL INTERCOM AMPLIFIER PERFORMANCE CHARACTERISTICS USING 2N342A TRANSISTORS





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Professional Group on Circuit Theory

One of the most fundamental subjects in the radio engineering profession is circuit theory. It forms a basic part of every engineer's schooling. The methods of analyzing the characteristics of circuits and the techniques of designing circuits for optimum performance find wide application in practically every branch of communications and electronics. It is not surprising, therefore, that a substantial portion of the technical literature is devoted to this important subject.

While circuit theory is broad in its applications, the formulation of these methods and techniques is a sharply defined and well-developed field in its own right, requiring specialized knowledge of a high degree. The IRE Professional Group on Circuit Theory is playing a key role in furthering the development of this field by making this specialized knowledge readily available to those who need it.

The Circuit Theory Group began operating in April, 1949 and since that time has become one of the largest and most active Groups in the IRE. Its most important activity has been the publication of its own technical periodical, called IRE TRANSACTIONS on Circuit Theory. Begun in 1952, the TRANSACTIONS is issued quarterly to some 6700 members as a part of their assessment. An interesting and valuable feature of TRANSACTIONS is that many of the issues are built around a special topic. Among the topics thus treated are Servo-mechanisms, Circuit Stability, Network Approximation, Nonlinear Filters, and Time-Variable Networks.

The Circuit Theory Group has been active also in sponsoring meetings and in organizing sessions at several national conferences. Numerous meetings are also held in over a dozen cities by local chapters of the Group. The net result has been to give engineers a golden opportunity to keep fully informed of the many significant developments in this important field.

W. R. G. Baker

Chairman, Professional Groups Committee

ESC delay lines take off with America's talking satellite

On December 18, 1958, the world entered a new era of communications with the successful orbiting of an Atlas ICBM—the Talking Satellite that broadcast President Dwight D. Eisenhower's Christmas message to the world. Circling the earth at a speed of more than 17,000 mph, the Talking Satellite repeated the President's message, erased it, and received and rebroadcast new messages in both voice and code.

ESC Corporation is justifiably proud that its delay lines were selected to aid in this electrifying triumph for America and her electronics industry. Especially designed by ESC, these delay lines were used in the timing sequence for propulsion, the guidance system and the telemetering system.

As America's largest manufacturer of custom-built and stock delay lines, ESC has continually met the responsibility of leadership by providing virtually every type of delay line needed by defense and industry. If you have a delay line problem, let ESC's design staff suggest a custom-built answer.

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Lumped-constant delay lines • Variable delay networks • Continuously variable delay lines • Pushbutton decade delay lines • Shift
Pulse-forming networks • Miniature plug-in encapsulated circuit assemblies

Professional Group Meetings

(Continued from page 76A)

Broadcasting

Cleveland—January 22

"The Ampex Video Tape Recorder" (Description & Application), S. Stadig, KYW-TV.

"The Ampex Video Tape Recorder" (Mechanical & Electrical Circuits), A. Hopwood, KYW-TV.

Philadelphia—February 5

"A New Vidicon Camera," J. R. Kingston, Philco Corp.

"Multiplexing FM Using the Sum and Difference System for Stereo Broadcasting," R. B. Walsh, Station WJBR.

Washington, D. C.—January 26

"The Voice of America," E. T. Martin, U. S. Information Agency.

Broadcast and Television Receivers

Los Angeles—January 22

"Compatible Full-Fidelity F.M. Multiplex System—Utilizing Narrow Band Sub-Carrier," H. N. Parker, Calbest.

Circuit Theory

Albuquerque-Los Alamos—September 24

Planning of general type of technical meeting to have during coming year, and fixing meeting dates.

Albuquerque-Los Alamos—November 24

"On the Use of the Scattering Matrix in Network Analysis," Lt. W. A. Whitaker.

Albuquerque-Los Alamos—January 26

"On the History of Circuit Synthesis and its Present Status," A. Melloh.

Los Angeles—December 9

"Transient Correction by Means of All-Pass Networks," J. C. Pinson, Autonetics, a Div. of North American Aviation.

"Matrices Relevant to Network Synthesis," P. Slepian, Hughes Aircraft Co.

Communications Systems

Oklahoma City—September 23

Paper originally presented to National Conference of Police Officers Communications at Baltimore, Md., August, 1958 by Castleberry and Hansen of G.E. Co., T. W. Stevens, G.E. Co.

Oklahoma City—November 27

"Precise Measurements of Broadcast Radio and Television Channel Frequencies, using General Radio Standard," E. Black, Yellow Cab Co. and Radio Station KOCY.

(Continued on page 86A)

NEW ECONOMY OF CIRCUIT DESIGN with new NPN power transistors

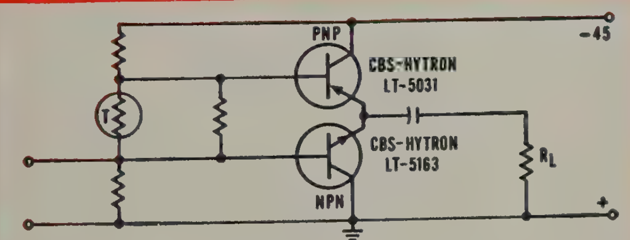
. . . in complementary push-pull circuits

These new NPN transistors in the complementary push-pull circuit eliminate input and output transformers with many resulting advantages:

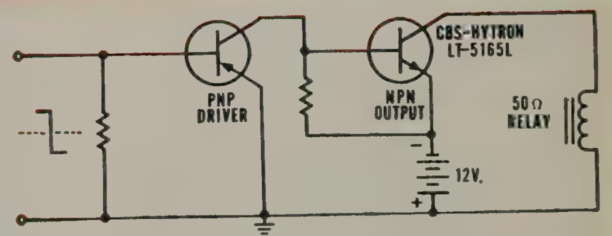
1. Economy. 2. Miniaturization. 3. Improved frequency response. 4. Ease of applying negative feedback.

. . . in complementary cascade circuits

CBS-Hytron NPN power transistors are also intended for single-ended cascade output circuits, and three types are designed for use with inductive loads such as output transformers, motors and relays. The simplified direct-coupled circuits permit economies through elimination of components.



COMPLEMENTARY CLASS B AMPLIFIER



DIRECT-COUPLED COMPLEMENTARY CONTROL AMPLIFIER

FIRST COMPREHENSIVE LINE OF NPN POWER TRANSISTORS

Type	Max. W Diss. (25°C Ambient*)	Max. V_{CBO}	Max. V_{CES}	Min. h_{FE} ($I_C = 0.5A$)	Max. Thermal Res. °C/W	Similar PNP Types
FOR RESISTIVE-LOAD CIRCUITS						
2N326	7	35	35†	30	8	LT-5025
LT-5165	12	35	30#	30	5	LT-5025
LT-5163	12	60	45#	30	5	LT-5031
LT-5164	12	80	60#	30	5	LT-5039
FOR INDUCTIVE-LOAD CIRCUITS						
LT-5165L	12	35	30φ	30	5	LT-5031
LT-5163L	12	60	45φ	30	5	LT-5039
LT-5164L	12	80	60φ	30	5	LT-5048

All seven types have: Max. collector current, 3 amps; storage temperature, −65 to +85°C; min. cutoff frequency, 150 kc. *Adequate heat sink required.
†I_{CES} = 1 ma. max. #I_{CES} = 10 ma. φI_{CES} = 300 ma.

New design economies in complementary push-pull and cascade circuits are made possible by these CBS-Hytron NPN power transistors. They feature high voltages . . . up to 80 volts. Their proven reliability (they meet the MIL-19500A specification) is what you might expect from specialists in reliable NPN switching and PNP power transistors. Check the typical circuits and abbreviated data for this versatile and comprehensive line. Write for complete technical Bulletin E-332 . . . and, if you wish, for Bulletin E-288 on CBS-Hytron PNP power transistors for complementary circuits.

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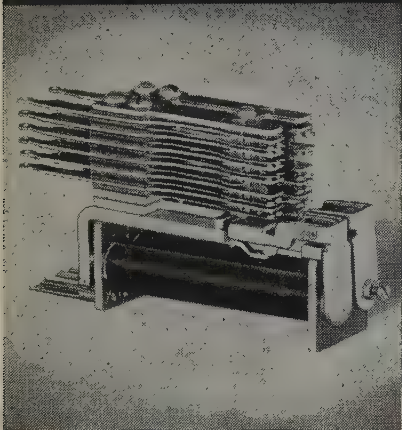
CBS-HYTRON, Semiconductor Operations
A Division of Columbia Broadcasting System, Inc.

Stromberg-Carlson "TELEPHONE QUALITY" Relays



. . . available immediately for any part of your operation that depends on electromechanical switching.

Proven by many years of meeting the exacting requirements of the telephone industry, these twin-contact relays of unsurpassed reliability are available in many types. The following are representative:

Type A: general-purpose relay with up to 20 Form "A" spring combinations. This relay is excellent for switching operations.

Type B: a gang-type relay with up to 60 Form "A" spring combinations. **Type BB** relay accommodates up to 100 Form "A" springs.

Type C (illustrated): two relays on the same frame. A "must" where space is at a premium.

Type E: has the characteristics of Type A relay, plus universal mounting arrangement. Interchangeable with many other makes.

Complete details and specifications on all Stromberg-Carlson relays are contained in our *new relay catalog*. Contents include: spring combinations, table of equivalents, contact data, variations and special features, plus complete mounting and cover information.

The catalog is available on request.

STROMBERG-CARLSON

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Telecommunication Industrial Sales
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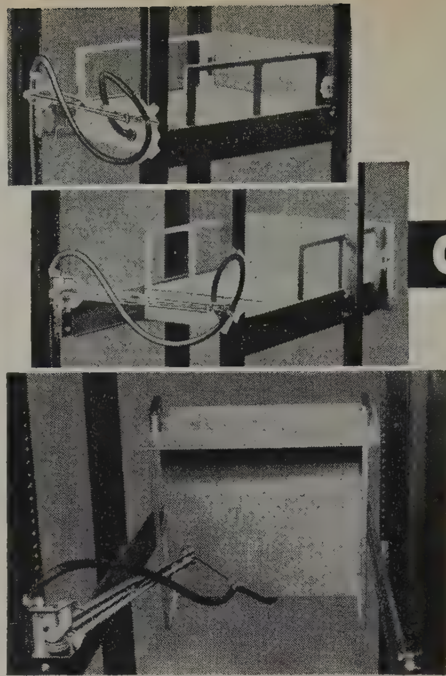
S-C



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CABLE RETRACTOR

Withdrawal of a chassis for service and its return to position no longer presents the old bugaboo of cable entanglement with and damage to tubes and components in the chassis immediately below it.

This new cable retractor's double action maintains a constant tension and correct suspension of cable at all times—permits adequate cable length for full extension and tilting of chassis without hazard of snagging.

May be used with all types of chassis or drawer slides, is adjustable to fit varying chassis lengths, is simple to install, and has proven thoroughly reliable in operation.

Mounts on rear support rails on standard 1¾" hole increments. Cadmium plated cold rolled steel.

CABLE RETRACTOR INSTALLED
One support rail is shown cut away to more clearly illustrate complete absence of cable sag at every stage.
TOP—installation with slide closed
MIDDLE—chassis partly withdrawn
BOTTOM—slide extended, chassis tilted

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ONE SOURCE... for VENTILATED RELAY RACK CABINETS, CONTROL CONSOLES, BLOWERS, CHASSIS, 'CHASSIS-TRAK'*, RELATED COMPONENTS

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. . . TINKERING TIME
. . . POOR PERFORMANCE
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Satisfactory design and profitable operation of high vacuum equipment require specialized knowledge and components. Even if you're already a vacuum expert, you can avoid lengthy study, tedious calculations, and costly pitfalls by asking for the assistance of NRC vacuum specialists. That's the quick, sure, no-cost way to assure that you benefit from the lessons learned in thousands of vacuum installations.

NRC Equipment Corporation is your one convenient source for service-proved components, equipment, and systems especially designed for high vacuum service, PLUS technical help in selecting the items which will best suit your own needs.

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1,068,111	99	0	ELECTRONICS, Oct. 1, 1957, pg. 167
5,460,000	600	1	ELECTRONICS, Oct. 1, 1957, pg. 167
1,250,000	125	0	PHILCO REPORT, Feb. 10, 1959
16,000,000	10,192	2	WJCC REPORT, Feb. 1957
8,640,000	8,000	2	PHILCO REPORT, Feb. 12, 1959
19,196,232	18,601	3	PHILCO REPORT, Nov. 19, 1958

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‡Failures due to all causes including human error.

*Documented service hours in these six computers only. Total transistors hours in similar circuits are many times this amount.

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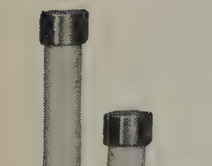
Write for Bulletin 301 with complete description and specification data. Lapp Insulator Co., Inc., Radio Specialties Division, 217 Sumner St., Le Roy, N. Y.



Professional Group Meetings

(Continued from page 82A)

Oklahoma City—January 27

"Transmission Lines and Antenna Matching," T. Wright, FAA.

Communication Systems/Vehicular Communications

Philadelphia—January 5

"A Review of Single-Side-Band Transmission Systems," E. A. Laport, RCA.

Philadelphia—January 6

"Automatic Ground/Air/Ground Communications System," R. E. Davis, RCA.

Component Parts

Baltimore—January 27

"Recovery Time Characteristics of Semi-Conductor Rectifiers used in Fast Switching or High Frequency Circuits," L. G. McPherson, Westinghouse Elec. Corp.

Buffalo-Niagara—October 22

"Optics in Electronics," E. Dixon, American Optical Co.

(Continued on page 90A)

X-Y RECORDING never before possible with electromechanical instruments can now be done with the new Sanborn Model 670 X-Y Recorder. Direct writing on ultraviolet-sensitive recording paper by a beam deflected by optical galvanometers makes possible the combination of fast writing speed and 130 cps frequency response not found in any other X-Y recorder. Transistor characteristics, acceleration and vibration of mechanical parts and events of similar short duration can be recorded with linearity of *1% of full-scale* and at trace speeds as fast as *2500 inches per second.* Square wave response exhibits no greater than ½% overshoot *at any amplitude;* sensitivities as high as 62.5 uv/inch (depending on preamplifier used).

PLOTS OCCUPY AN 8″ x 8″ RECORDING AREA and can be previewed or monitored on the instrument's phosphorescent screen. An Axis Record switch to print X and Y axes on the record, and a Beam Intensity Control to assure maximum trace clarity, are among the front panel controls provided. An 8″ x 8″ sheet of the ultraviolet-sensitive chart paper (stored in drawer at base of cabinet) is easily placed on the back of the hinged screen. Brief post exposure in normal room light is the only developing process.

OPTIONAL INTERCHANGEABLE PREAMPLIFIERS for each axis presently include the Model 850-1300B DC Coupling and Model 850-1200 Phase Sensitive Demodulator; a Carrier Preamplifier, High Gain Preamplifier and a time base generator are now in development. Driver Amplifiers are compact, fully transistorized plug-in units with single-ended input and output. Galvanometers are low resistance, low voltage units of rugged, enclosed construction; sensitivity and damping are independent of coil temperature. Accessible, unitized circuitry also extends to the power supplies—a front-panel plug-in for both preamplifiers and a second supply for both driver amplifiers. A built-in blower provides constant, forced filtered air cooling. The Recorder can be rack mounted in 15¾″ of panel space, or housed in its own 20″ x 20″ x 21¼″ optional portable cabinet.

Ask your local Sanborn Sales-Engineering Representative for complete information on the Model 670 X-Y Recorder, or write the Industrial Division in Waltham, Mass.

SANBORN COMPANY

INDUSTRIAL DIVISION • 175 Wyman Street, Waltham 54, Massachusetts

Professional Group Meetings

(Continued from page 86A)

Philadelphia—January 6

"Progress on International Standards," L. Podolsky, Sprague Elec. Co.

Electron Devices

Boston—December 10

"Wide-Band, Multicavity Klystrons," L. D. Smullin, M.I.T.

Boston—January 16

"Status of Transistor Technology," W. J. Pietenpol, Sylvania Electric Prods.

Los Angeles—January 19

"Panel Light Amplifiers for Intensification and Storage of Images," B. Kazan, Hughes Aircraft Co.

Washington, D. C.—January 19

"Field Emission," W. P. Dyke, Linfield Res. Inst.

Electron Devices/Microwave Theory and Techniques

San Francisco—October 15

"Nonlinear-Reactances, Their Characteristics and Properties," B. Salsburg, Airborne Instruments Lab.

San Francisco—October 22

"Back-Biased Diode Parametric Amplifiers," H. H. Heffner, Stanford Univ.

San Francisco—October 29

"Large Signal Properties of Parametric Amplifiers," K. Kotzebue.

San Francisco—November 12

"Ferromagnetic Parametric Amplifiers," P. H. Vartanian, Microwave Engineering Labs.

Electronic Computers

Baltimore—November 18

Field trip demonstration of Missile Master Center located at Fort George Meade.

Baltimore—February 10

"A High Speed Analog to Digital Converter," J. M. Bentley, Westinghouse Elec. Corp.

"The BRL Computer," J. G. Gregory, Ballistics Res. Lab.

Detroit—November 5

"Application of Computer Techniques to Machine Control," J. L. McKelvie, Bendix Research Lab.; H. H. Schatz, Bendix Industrial Controls Sec.; F. E. Booth, Bendix Industrial Sec.

(Continued on page 100A)

C TYPE MAGNETS in a wide range of sizes to meet your design needs in ★ Transverse Field Isolators ★ Differential Phase Shifters ★ Duplexers

Arnold C-type Alnico Magnets are available in a wide selection of gap densities ranging from 1,000 to over 7,500 gausses. There are six different basic configurations with a wide range of stock sizes in each group.

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SEMICONDUCTORS

Edited by

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Bell Telephone Laboratories
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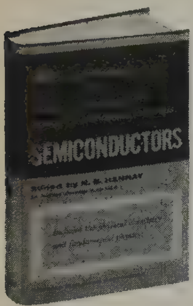

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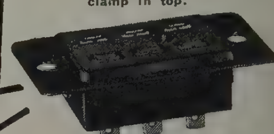

S-2406-SB Socket with shallow bracket for flush mounting.

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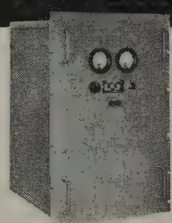
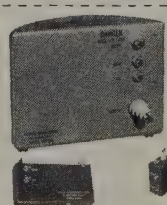


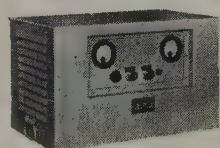
MISSILE-GROUND

Perkin has developed Tubeless Magnetic Amplifier Regulated DC Power Supplies for missile launching and check-out, with ratings (@ 24-32 volts) of 30 to 500 amperes: with provisions for rack panel mounting and remote sensing. Perkin ground power supplies are now being used in the *Thor, Atlas, Bomarc, Vanguard* and other missile programs.



RADAR-AIRBORNE

There are more than 6,000 Perkin units operating in Military and Commercial Radar Systems. Ratings from 6 to 12KV are available. Nominal AC input frequency is 400 cps; Temperature Range—up to +125°C and above. Typical Weight: 2 lbs. No tubes, moving parts or vibrating contacts; epoxy resin potted.



RADAR-GROUND

Hundreds of Perkin Ground Radar Systems are in operation throughout the country. Built to MIL-E-4158. Typical unit is rated at 24-32 volts @ 100 amperes, ±½% regulation and 1% rms ripple. All units conservatively designed.

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PERKIN ENGINEERING CORPORATION
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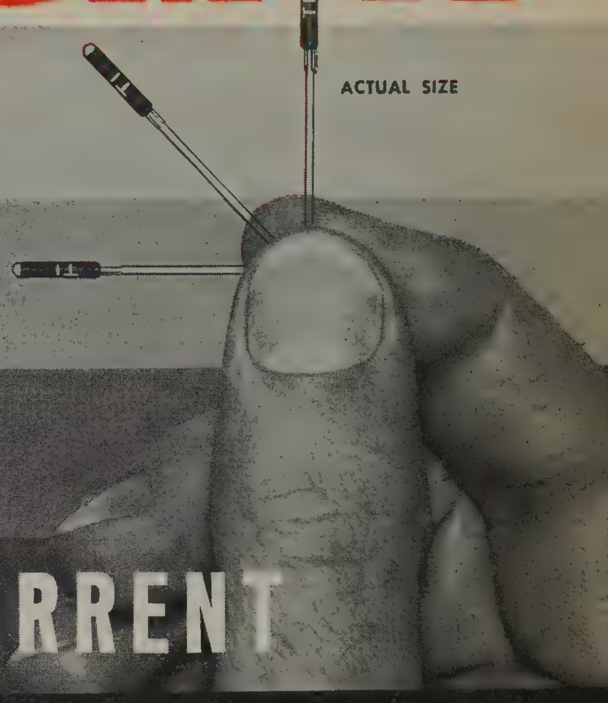
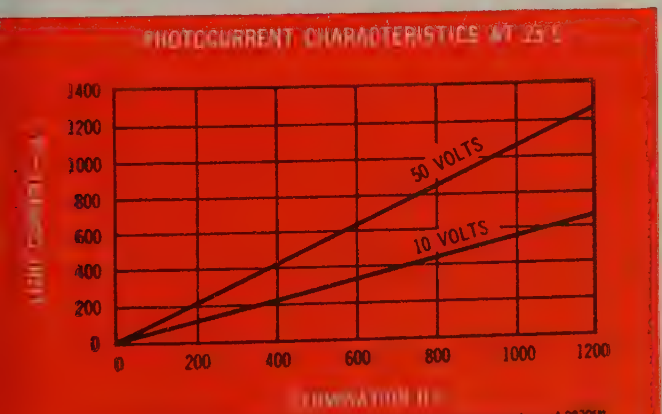
1N2175 SILICON PHOTO-DEVICE SWITCHES FROM DARK TO LIGHT CURRENT IN 2 MICROSECONDS



TI 1N2175 subminiature unit is ideal for punch-card or tape data processing, and many other control systems



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Easily activated, the 1N2175 switches from a low dark current of only 0.5 μa to a high light current of 1200 μa at 1200 ft-candles — *within 2 μsecs.* Rated at 250 mw at 25°C, the 1N2175 operates over a range of 1-50 volts, and derates linearly to 125°C. Minimum operating temperature is −55°C.

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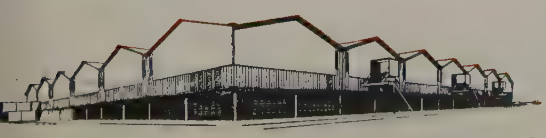

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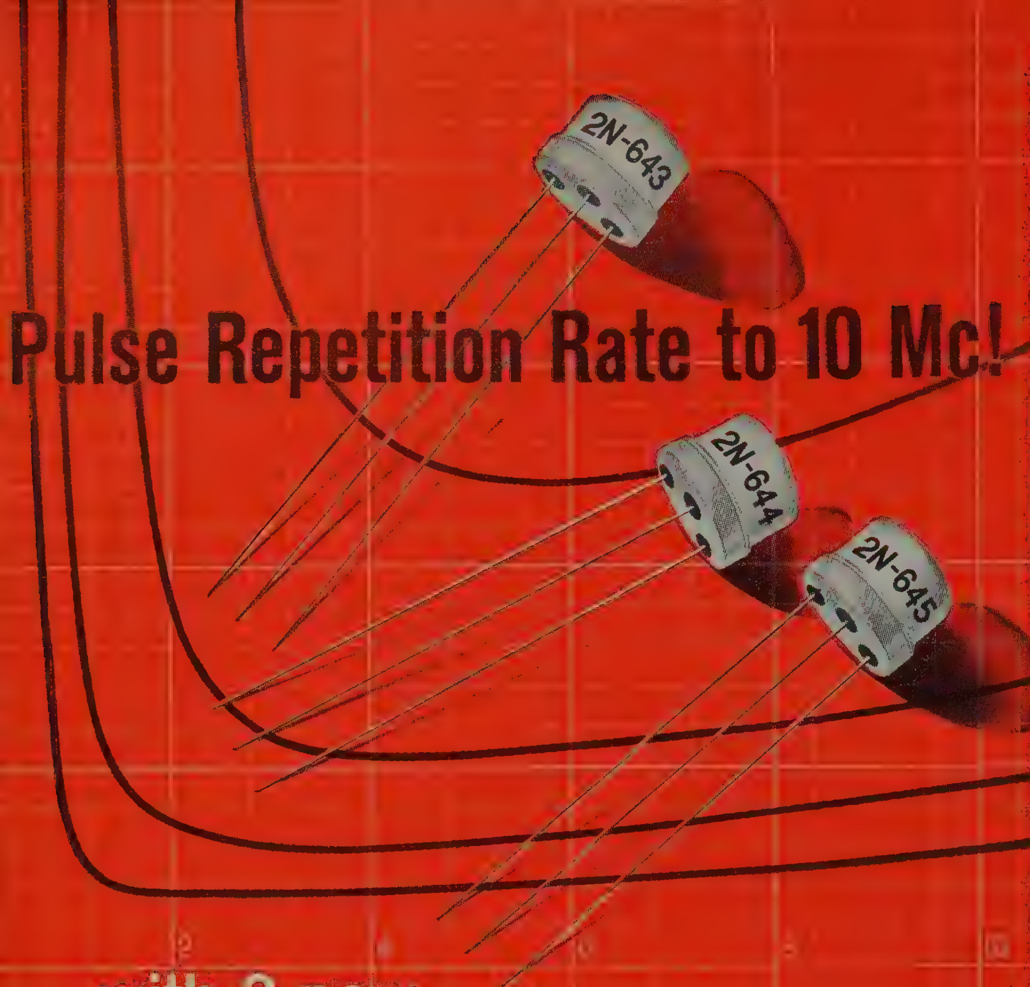

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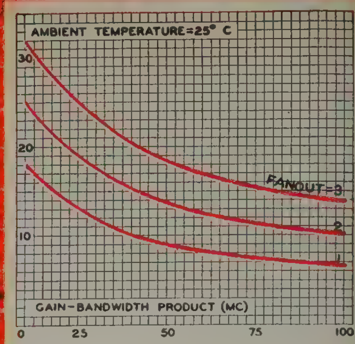
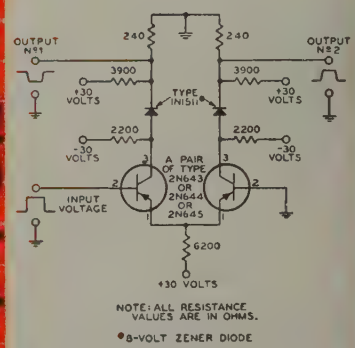


Curves illustrate typical delay time per stage vs. gain-bandwidth product and fanout for the switching circuit shown below.



TYPE	2N643	2N644	2N645
Minimum gain-bandwidth product* Mc	20	40	60
Minimum collector** breakdown volts	30	30	30
Minimum DC current transfer ratio*	20	20	20
Maximum collector capacitance µµf	5	5	5

*Collector Volts = -7, collector ma = -5
**Collector Current = 100 µa

RCA-2N643, RCA-2N644, and RCA-2N645 feature controlled minimum gain-bandwidth products, of 20, 40, and 60 Mc

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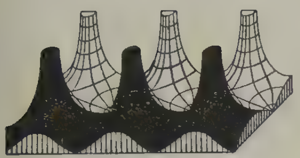
April, 1959
Vol. 47 No. 4

Proceedings of the IRE

Poles and Zeros



Roses To Ye Editors. An important adjunct of electronic and IRE communications, not too generally appreciated, is the local IRE Section publication. Now the products of 47 Sections or almost half of all, these magazines have come fully of age as a means of passing Section, Professional Group Chapter, and national news to IRE members. From the six or eight-page publication of the small Section to the 32 to 40-page output of the large, they all provide a means of informing the members of Section and PG Chapter meetings and programs, and many publish abstracts of past programs for those members who unfortunately missed the most recent meeting.

With total meetings mounting into the range of 12 to 20 in some Sections, augmented by special symposia, technical lecture series, or by Regional activities, the superiority of the Section magazine over smoke signals, carrier pigeons, post cards, or any earlier technique need not be argued. What has not been so apparent is the excellent editorial work being done by our volunteer Section editors. The addition of editorials, Section personnel and business news, ladies' columns (to placate the wife for those IRE nights out), student branch or local engineering college news, even prize contests, has made what could have been dusty dissertations into praiseworthy prose.

These editors have also learned well that basic rule of local journalism—"names and pictures of local faces, will put the paper through its daily paces," and embellish their columns with pictures of authors, local Fellow awardees, Section officers and other personalities of local interest. This often also permits noncommercial bows to the supporters of it all, the local and national advertisers, and disseminates from their printed pages the feeling that the Section is a very well integrated technical and social family.

Much originality has gone into the titles of these papers, and we cite a few as "The Missile" (Alamogordo-Holloman), "Crosstalk" (Detroit), "The Long Island Pulse," or the "Kansas City Local Oscillator." If Albuquerque-Los Alamos did not already possess a good title in "The Blast" we might suggest "The *Q* Point," and we feel that the Washington Section overlooked one in the possibilities in "The DC Amplifier." In fact, always desiring to be helpful, this Editor offers free and with no rights reserved, the following titles to a few Sections not yet having publications. Why do we not have: "The Rochester Radiator," "The Beaumont Beeper," the Buffalo "BISONic News," "The Twin Cities Stereo," or "Phoenix Phonics?"

Pills and Pulses. As you scan the photographs and biographies of the new IRE Fellows in this issue, you will not want to miss an historic first—Dr. Lee Lusted who is the first medical doctor to be elected an IRE Fellow. This gives formal and significant recognition to the growing bond between the physician, the physiologist, the biologist, and the electronic engineer or the physical scientist.

Many of us at present may think of this relationship as somewhat of a one-way street in which electronic measuring or recording techniques are used to obtain medical information. However, this may soon be replaced with a two-lane boulevard in which knowledge from medicine and physiology will return to aid us in the solution of future problems. Especially is this true in the areas of communication, memory, and navigation, where knowledge of the human brain, our sensory organs, and the navigation methods of birds, bats, and fish may be needed before we become really sophisticated in those areas. Needless to say, our Professional Group on Medical Electronics in which Dr. Lusted has been active, is a strong motivating force in this exchange of knowledge.

It should also be mentioned that another new Fellow, Herman P. Schwan, is the third biophysicist to be so honored, and that Dr. Alfred Goldsmith, our Editor Emeritus, has long been a member of several noted medical societies. Thus do the barriers between fields of science continue to fall.

A Reminder. Those of you not attending the March Convention may have missed the announcement that IRE Convention Record parts will no longer be distributed free to PG members. After study by the Editorial Board, the PG committee, and the Board of Directors, this action was taken to encourage broader publication of selected Convention papers by the PG Transactions—which are then available to the PG member. The move also permits a drastic reduction in price for those completing their files by purchase. Convention Record orders must be placed with Headquarters before April 30. Information on how to order and prices for the various parts appear on page 20A of this issue.

This And That. Attention should be turned to a letter in this month's Correspondence Section (p. 584) from Mr. Floyd W. Hough, chairman of an American Geophysical Union committee for Study of the Metric System in the United States. He asks for expressions of interest in a proposal to adopt the metric system in the United States through a phased procedure covering 33 years, or a generation. With the MKS system already well established in our area, we might then be able to forget that a meter is 39.37 inches, that 32 ounces made up a quart but 16 ounces make a pound—or that sometimes there are five quarts in a gallon.—J.D.R.

Bernard M. Oliver

Director, 1959–1961

Bernard M. Oliver (A'40–M'46–SM'53–F'54) was born on May 27, 1916, in Santa Cruz, California. He was graduated from Stanford University, Stanford, Calif., with the B.A. degree in electrical engineering in 1935, and in 1936 received the M.S. degree from The California Institute of Technology in Pasadena. The next year he studied in Germany as an exchange student under the auspices of the Institute of International Education. At the age of twenty-three he received the Ph.D. degree, magna cum laude, in electrical engineering from the California Institute of Technology.

Dr. Oliver was with the Bell Telephone Laboratories in New York from 1940 to 1952, where he worked on the development of automatic tracking radar, television, information theory, and high efficiency systems. In 1952 he joined the Hewlett-Packard Co., Palo Alto, Calif., as director of research, and in 1957 became vice-president in charge of research and development there. For the past several years he has also been a lecturer in electrical engineering at Stanford University.

Dr. Oliver holds twenty-seven U. S. patents in the field of electronics; others are pending. He has also contributed several papers to the PROCEEDINGS of the IRE.

He has served as Chairman of the IRE San Francisco Section, and for several years was Vice-Chairman of WESCON

Scanning the Issue

A New Concept in Computing (Wigington, p. 516)—The late John Von Neumann, who contributed so much to so many fields, is especially renowned for the leading role he played in the birth of the modern electronic computing machine. One of his last major contributions was contained in a patent submitted in 1954 and granted posthumously only 16 months ago. This paper presents a description and explanation of the new computing scheme disclosed in that patent. In it Von Neumann proposed that digital information be represented by the *phase* of a sine wave. He suggested that this could be done by using a nonlinear reactance device as a subharmonic generator to produce oscillations at one-nth the pumping frequency. It can be shown that the subharmonic wave will have a choice of n instants during the period of each cycle when it can start off in phase with the pumping wave. In other words, the subharmonic wave will lock into any of n phases. The choice of phase can be controlled by the phase of an input signal at the subharmonic frequency. This can be thought of as a mechanism for producing n logic states. When several subharmonic generators are connected in a circuit, the output of one will change the state (phase) of another, producing an n-ary logic system. This idea is closely related to the parametron, which was independently developed in Japan and which has aroused so much interest in the past couple of years. The parametron uses nonlinear oscillators operating in the one-megacycle range. However, it is the possibility of applying this scheme to microwave frequencies that makes it of such outstanding importance, because microwaves would greatly increase the operating speed of computers. This work will be of great importance not only to computer engineers but also to all those who are interested in parametric amplifiers.

Stored Charge Method of Transistor Base Transit Analysis (Varnerin, p. 523)—The operating frequency of a transistor can be increased by reducing the base transit time, presumably either by increasing the velocity of the carrier flow or by reducing the base width. Velocities can be increased by grading the base impurity densities to produce built-in aiding electric fields. This principle has been widely used in recent high-frequency designs since the advent of diffusion techniques for fabricating transistors. Base transit time, as presently defined, is primarily a mathematical concept; that is, it is expressed in terms which have mathematical validity but which do not all have a direct physical interpretation. In this paper, the author re-expresses base transit time in terms of stored charge per unit emitter current—terms which are conceptually much simpler to deal with and which greatly facilitate the understanding of high-frequency performance. Using this fresh viewpoint, the author readily shows that base thickness is more important than the built-in field in determining base transit time. In fact, this gives rise to a seeming paradox: shorter transit times are possible with retarding than with aiding fields, because they permit a narrower base width. Both the new method of analysis and the results will be of wide interest to the considerable body of engineers interested in high-frequency transistor design.

The Hall Effect Circulator—A Passive Transmission Device (Grubbs, p. 528)—One of the notable features of the 1950's has been the variety of devices that have been developed which exhibit nonreciprocal transmission characteristics. These developments have led not only to new components and improved systems but to new methods of network analysis as well. The development of ferrites has provided nonreciprocal devices for the microwave range, while the exploitation of the Hall effect in semiconductors is leading to devices with similar functions for use at the lower "wire" frequencies. Hall effect gyrators and isolators have already been investigated for use in wire circuits. This paper describes an important addition to the nonreciprocal device family, a Hall effect circulator.

Theory of the Crestatron: A Forward-Wave Amplifier (Rowe, p. 536)—Past experimental data have indicated that gain can occur in a traveling-wave tube even though the voltage is so high that a growing wave cannot exist, but the phenomenon has not heretofore been understood. The author has now developed a theory which satisfactorily explains the previous data. To verify his work, he has built a tube which utilizes this new mode of operation. The tube proves to have moderate gain, good efficiency and, perhaps the most significant feature, a very short length, which suggests the interesting possibility that the device might operate with little or no magnetic focusing field.

Theory and Experiments on Shot Noise in Silicon *P-N* Junction Diodes and Transistors (Schneider and Strutt, p. 546)—It has been found that the theory which describes the noise performance of germanium diodes and transistors does not hold true for silicon devices. This is not surprising since the electrical characteristics of silicon devices differ to some extent from those of the corresponding germanium types. In fact, it was not until a year and a half ago that this difference in the characteristics of silicon junctions was adequately explained. Armed with this information, the authors have succeeded in deriving new noise expressions which are valid for silicon devices operating at low current densities. This paper is of great interest in understanding the noise behavior of silicon devices—indeed, it is the first paper to tackle the problem.

A Constant-Temperature-Operation Hot-Wire Anemometer (Janssen, *et al.*, p. 555)—This paper discusses a type of instrument that is very useful in studying turbulent air flow. When a small, electrically heated wire is placed in the air flow, turbulence will cause variations in the cooling effect of the air stream on the heated wire. These temperature changes, in turn, cause the resistance of the wire to vary. By incorporating the heated wire in an appropriate circuit—in this case, a dc bridge and a dc differential amplifier—air velocity fluctuations containing frequencies of several tens of kilocycles can be measured electrically. While the instrument is not new, the application of a chopper stabilized amplifier described here is novel, as is the use of signal flow graphs in an analysis of this sort.

IRE Standards on Waveguide and Waveguide Component Measurements (p. 568)—This standard provides very useful and important information on the proper methods of measuring some two dozen major quantities which characterize waveguides, waveguide components, and the associated electromagnetic fields.

IRE Award Winners (p. 598)—Last month at the annual IRE Convention banquet in New York, over 1000 persons crowded into the Grand Ballroom of the Waldorf-Astoria Hotel to witness what in many respects is the climactic event of the year. On that occasion the leading contributors to the progress of our profession were singled out for recognition by the IRE—seven by means of IRE's highest awards and 76 by elevation to Fellow grade. The photographs and award citations of this distinguished group are presented in a special section of this issue.

IRE Transactions Index (follows p. 628)—During 1958 the Professional Groups of the IRE published 81 issues of Transactions, comprising more than 800 papers and letters. So substantial has the Transactions activity become that it now represents half of the total technical publication output of the IRE. This wealth of material has been catalogued by tables of contents, by authors, and by subjects in the 1958 index which appears toward the rear of this issue.

Scanning the Transactions appears on page 604.

A New Concept in Computing*

R. L. WIGINGTON†, MEMBER, IRE

Summary—**A new computing scheme was proposed by von Neumann in a patent[1] submitted in 1954 and granted posthumously last December. This paper is an explanatory statement of those ideas. The concept of using the phase of a sine-wave signal as an information-bearing medium which together with majority logic permits the realization of logic operations is described in detail. Simple logical aggregates are given as examples.**

Introduction

Von Neumann recognized the limitations of computing speeds inherent in the existing technology due to device operation times, signal propagation delays, and transmission distortion of information video pulses. Of course, we cannot reconstruct the thinking that led him to the proposed solution, but the attractiveness of large bandwidths which could be obtained at microwave frequencies and of representation of digital information by distinct phases of an RF signal (neither of which had been exploited in computer technology) no doubt seemed fertile ground for investigation. Whatever the prompting force was, von Neumann proposed a computing scheme using RF techniques which is potentially faster if employed at microwave frequencies than present conventional methods. The same methods will also work at lower frequencies.

The availability of an element of the following nature was assumed:

a) An element is available that has both L and C, one of which is nonlinear.
b) The dissipation in the element is not great in comparison to the nonlinearity. This dissipation may be resistive loss or a hysteresis.
c) The element has approximately linear operation for small signals having a resonant frequency (which later is referred to as f_0).

Nonlinear Reactance Analysis

To see that such an element can exist, consider the nonlinear capacitor that has no loss and no hysteresis[2] and is mounted in multiply-tuned circuits. The bare outline of the analysis will be presented. Full discussion of it has been published by Manley and Rowe.[3] As pointed out in that reference, consideration of nonlinear reactances as elements to transfer power from one frequency to another is not a new subject, having been discussed by Hartley in 1916.

Consider the circuit in Fig. 1.

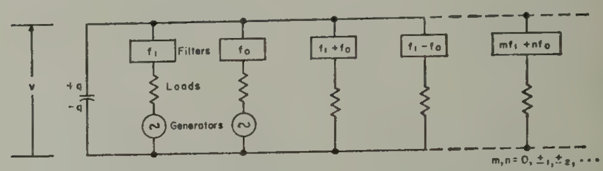


Fig. 1—Generalized nonlinear reactance circuit.

$v=f(q)$ is an arbitrary, single-valued function, being in general nonlinear.

The filters have zero impedance at the labeled frequency and infinite impedance otherwise.

Such a circuit might be a number of tuned circuits in parallel with the nonlinear reactance (which also may be inductance rather than capacitance). Two of the tuned circuits contain RF generators.

The charge on the capacitor, the current into it, and the voltage across it may all be expressed as double Fourier series in the frequencies mf_1+nf_0 for $m, =0, \pm 1, \pm 2, \cdots$. By manipulation of these expressions[4] one may derive others giving the summation of real power at all frequencies into the nonlinear element in terms of double integrals over complete cycles of f_1 and f_0 wherein the integrand is $v=f(q)$.

The variable q, hence $f(q)$, is periodic in f_1 and f_0 and thus for an element without hysteresis these double integrals are identically zero. From this are obtained the Manley-Rowe conditions, namely:

$$\sum_{m=0}^{\infty}\sum_{n=-\infty}^{\infty}\frac{mW_{m,n}}{mf_1+nf_0}=0 \qquad m, n = \text{integers}$$

$$\sum_{m=-\infty}^{\infty}\sum_{n=0}^{\infty}\frac{nW_{m,n}}{mf_1+nf_0}=0$$

where $W_{m,n}$ is the real power at frequency mf_1+nf_0 into the nonlinear reactance.

Of course, a reactance without hysteresis, whether linear or nonlinear, can dissipate no energy. Therefore, as must be true and as can be obtained from the above conditions, the summation of energy at all frequencies into the reactance is zero, and power supplied from one source at a given frequency will show up in another branch of the circuit at another frequency. The exact manner in which this takes place is controlled by the above conditions.

* Original manuscript received by the IRE, October 15, 1958; revised manuscript received, December 31, 1958.

† National Security Agency, Dept. of Defense, Ft. Meade, Md.

[1] J. von Neumann, "Non-linear Capacitance or Inductance Switching, Amplifying, and Memory Organs," U. S. Patent No. 2,815,488, issued December 3, 1957, assigned to the IBM Corporation.

[2] These restrictions may be relaxed somewhat in the practical case, but are assumed here to make explanation simpler.

[3] J. M. Manley and R. E. Rowe, "Some general properties of nonlinear elements—Part I. General energy relations," Proc. IRE, vol. 44, pp. 904–913; July, 1956.

[4] Complete discussion is given in Manley and Rowe, *ibid.*

The circuit initially pictured was a general circuit; variations of it may lead to modulators, demodulators, amplifiers, oscillators, or combinations of them. Of interest to this discussion is a subharmonic generator, as shown in the circuit of Fig. 2. V_G is an RF generator, essentially an ac power supply. The f_1 filter and R_G correspond to the resonant tank of the RF generator. The f_0 filter and R_L are the load tank at the subharmonic frequency desired. The filters at all other frequencies are terminated in pure reactances, X, which for simplicity may be open or short circuits. There will be no power loss at these other frequencies.

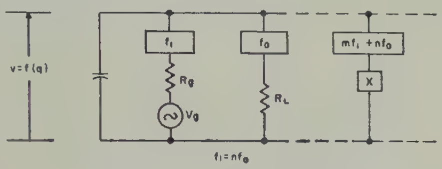


Fig. 2—Generalized subharmonic generator.

The only terms of interest in the Manley-Rowe conditions are for $(m, n) = (0, 1)$ and $(0, n)$. The rest of the $W_{m,n}$ are zero. This gives

$$\frac{W_{0,1}}{f_0} + \frac{nW_{0,n}}{nf_0} = 0$$

or simply

$$W_{0,1} = - W_{0,n}.$$

The minus sign indicates that all the energy put into the nonlinear reactance at the $f_1 = nf_0$ frequency is transferred to the load tuned to the f_0 frequency. This is a harmonic generator with an ideal efficiency of 100 per cent, assuming no losses in the reactive termination of the other frequencies. Of course, with real elements some losses, and resultant reduction of efficiency, will occur, but the basic (sub)harmonic generation process, unlike that of Class C multipliers or crystal multipliers, is not limited in efficiency.

Nature of Nonlinearity Required

It remains to show that an element with a threshold of subharmonic generation is possible. The transfer function for such an element is shown in Fig. 3.

The reactance function shown in Fig. 4 is, of course, idealized. In the actual case only a smoothly varying reactance would be obtainable which could be approximated by the lines in Fig. 4. At V_c, the nonlinearity is great enough to produce sufficient negative resistance at the subharmonic frequency to overcome passive circuit losses. At this level of the RF power supply, the circuit would break into oscillation at the subharmonic frequency and the threshold action be obtained. For V_{in} V_c, the amount of signal $V_{in} - V_c$ will be effective in producing subharmonic voltage. Whether this is a straight-line function or one that saturates will depend on other amplitude nonlinearities in the circuit.

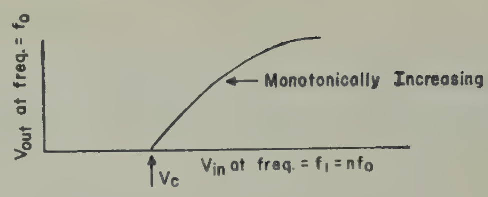


Fig. 3—Transfer function for subharmonic generator with threshold.

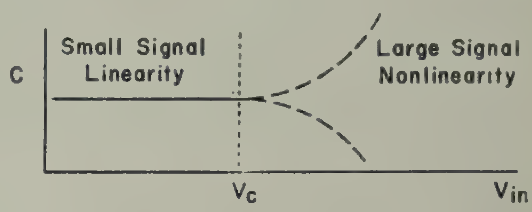


Fig. 4—Reactance variation to give threshold action.

Devices

Various solid-state elements have been proposed to perform the function of the nonlinear reactance.[5] An equivalent circuit for a nonlinear capacitance, as realized with a semiconductor diode, is shown in Fig. 5.

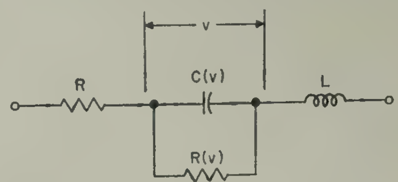


Fig. 5—Equivalent circuit for nonrectifying region of a semiconductor diode.

v = barrier voltage
R = bulk resistance
$R(v)$ = barrier resistance (which may be nonlinear)
L = lead inductance
$C(v)$ = barrier capacity (nonlinear).

The barrier capacity is the primary nonlinearity of the device. Among these components the following relationships are assumed:

$$\left.\begin{array}{l} R(v) \gg R \\ R_0 = \sqrt{L/C_s} \gg R, \end{array}\right\} \text{(The element has a reasonable } Q.)$$

C_s = small signal $C(v)$

$f_0 = 1/(2\pi\sqrt{LC_s})$

(Tuned to the subharmonic frequency desired.)

[5] A. Uhlir, Jr., "Two-terminal *p-n* junction devices for frequency conversion and computation," Proc. IRE, vol. 44, pp. 1183–1191; September, 1956.
H. Suhl, "The theory of the ferromagnetic microwave amplifier," *J. Appl. Phys.*, pp. 1225–1236; November, 1957.
F. Dill, Jr. and L. Depian, "Semiconductor capacitance amplifier," 1956 IRE Convention Record, pt. 3, pp. 172–174.
A. E. Bakenowski, "Small-Signal Measurements on Planar *P-N* Junction Diodes," "Task 8 Report on Crystal Rectifiers," Bell Tel. Labs., April 15, 1955.
A. Uhlir, "Frequency Conversion in *P-N* Junction Devices," "Task 8 Report on Crystal Rectifiers," Bell Tel. Labs., January 15, 1955.

This equivalent circuit as drawn includes both linear and nonlinear reactances. These may be physically separated in real devices. Research is being done at many laboratories on other practical means of achieving subharmonic response.

Information in Terms of Phase

The phase of the f_0 signal is determined by the phase of the f_1 signal. Qualitatively, this may be understood by observing that the oscillations in the f_0 circuit are not like the oscillations in an ordinary negative resistance oscillator, in which the power supply is dc and the phase of the oscillation is determined by noise when the oscillator is turned on. The nonlinear-reactance subharmonic generator is more akin to a crystal harmonic generator in which power is transferred from one frequency to another by a nonlinear element. This is true even though the action of the nonlinear-reactance device in general may be explained in terms of the apparent negative resistance reflected into each appropriate branch of the circuit.

Eliminating any constant phase shift between the fundamental and the nth harmonic, the phase relationships are examined more closely in Fig. 6.

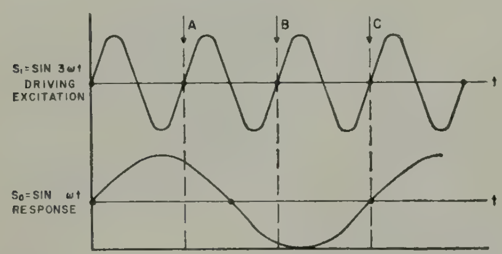


Fig. 6—Demonstration of three indeterminate phase relationships for the case $n=3$.

In the example in which $n=3$, the two signals are assumed in phase at $t=0$. (The $n=3$ example is chosen deliberately to point out important relationships and to avoid the $n=2$ case, in which simplicity may cause some essential points to be overlooked.) One full cycle of S_1 later, at point A, S_1 is the same as it was at $t=0$, but S_0 has a relative phase shift of $2\pi/3$. Similarly, at point B, S_1 is the same again, but S_0 now has a phase shift of $4\pi/3$. At point C, S_0 has a phase shift of $6\pi/3=2\pi$ or, like S_1, the same relative phase as when $t=0$. Thus there are, for $n=3$, three indeterminate phase relationships between S_0 and S_1, because in the continuous wave, S_1, one cycle is just like another. In the general case there are n such indeterminate relationships between the RF power supply (S_1) and the induced subharmonic response (S_0), each of which can represent a logical state.

Phase Selection and Control

If the power supply (S_1) is increased from zero amplitude past the critical amplitude, V_c, a subharmonic response S_0 will appear when the amplitude of S_1 is V_c. Since which of the n possible phases of S_0 will result is indeterminate, noise, or—indifferently—a small S_0 signal from another element, will dictate which phase will appear. Once the harmonic response has started, however, no external signal of less magnitude than $|S_0|$ can change the phase of the response. A simple cycle can be diagrammed (Fig. 7), plotting envelope amplitudes only. The S_0' signal would have no effect if present at any other time than when $|S_1|=V_c$ and is increasing. Note that

$\|S_0\|>S_0'$	(Amplification)
Duration of $S_0>$ Duration of S_0'	(Memory)
Phase of $S_0=$ Phase of S_0'	(Control or Toggle Action).

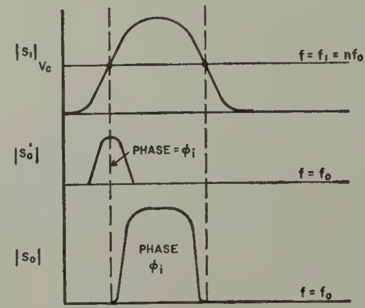


Fig. 7—Excitation, S_0', and response, S_0, during one cycle of pump modulation, S_1.

For practical realization of any computing scheme the electrical process must have natural or built-in margins. Noise, stray signal pickup, slight misadjustment of circuits, component changes with age, all can cause a certain malfunction of the equipment, if the parameters which determine the action of the machine must have precise values to cause action. For action must occur if the appropriate parameter falls within a certain region or margin, about the ideal value.

The controlling signal, S_0', has up to now been assumed to coincide exactly with one of the n possible phases, Φ_i, of the response, S_0, and to cause S_0 to respond with that phase, Φ_i.

In Fig. 8, the phase of S_0' is shown as being closer to the phase Φ_0 than either Φ_1 or Φ_2. The system will stabilize with S_0 at phase Φ_0, since this requires the minimum amount of energy in the presence of the S_0' signal to respond at the Φ_0 phase as compared with the other possible phases. The figure is not intended to illustrate the exact duration (in terms of number of cycles) of the transient condition, but is only a qualitative picture. The duration of the transient will depend on the rate of build-up of the S_1 envelope and the ratio of the response S_0 to the control S_0' during and after the start of the response. The magnitude of $\Delta\Phi$ and the effective damping constant of the transient condition will also be important.

One may conclude from this that the phase-locking property of the subharmonic generator has natural margins, and hence is a practical method.

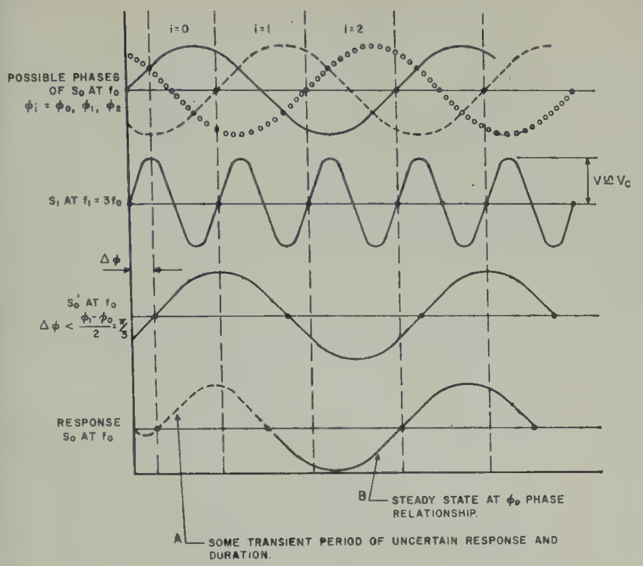


Fig. 8—Illustration of the natural margins of the phase locking mechanism.

Aggregates of Elements

In order to control the flow of information in a logical machine, ways must be devised to establish an order of control. For example, if information is to flow from A to B, then A must control B but B must not control A. Ordinarily this is no problem, because the nonreciprocal devices normally used in a logical circuit to provide gain or gating perform this function automatically, but in this case one must consider the control problem separately.

It is now necessary to devise a diagrammatic model of devices and their interconnection (see Fig. 9). The following symbolism and nomenclature will be used:

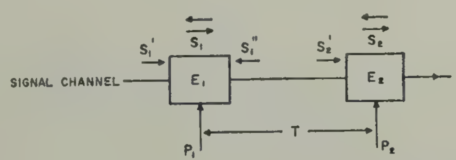


Fig. 9—Diagrammatic model of devices and their interconnection.

E_1, E_2 are elements such as have been described.
P_1, P_2 are amplitude-modulated AC power supplies.
S_1, S_2 are the output signals of E_1, E_2, respectively.
S_1', S_2' are the controlling signals for E_1, E_2, respectively.
S_1'' is the signal reaching E_1 from E_2.

The signal channel is an electromagnetic propagation path which permits propagation in either direction. The response S_1, from E_1 travels in both directions on the signal channel. The problem is to make E_1 control E_2 and not the reverse. Let T be the time delay of the signal between elements E_1 and E_2. Consider next the timing sequence in Fig. 10.

With the timing cycle as pictured, E_1 can control E_2 but E_2 cannot control E_1. If the relative timing (*i.e.*, the

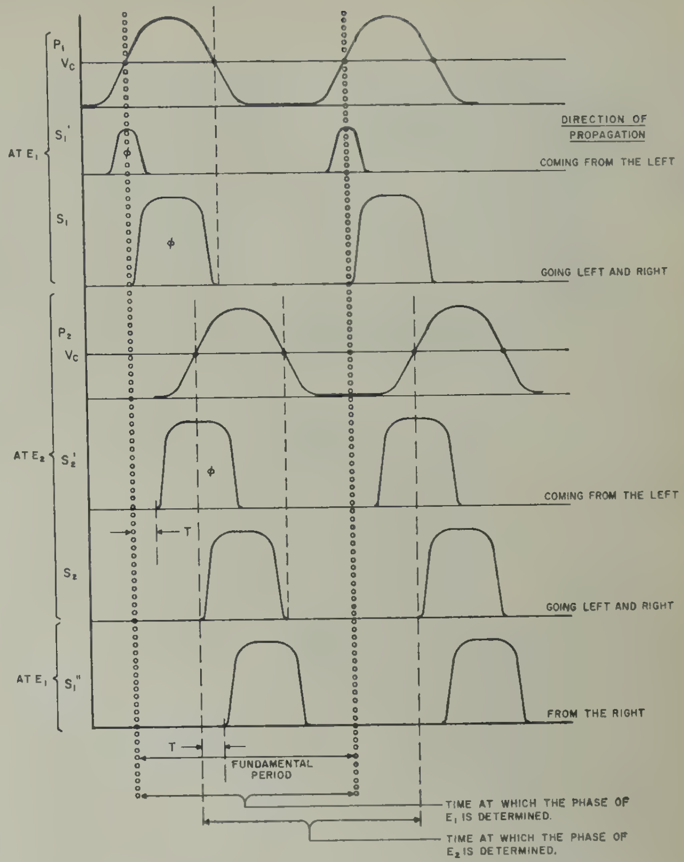


Fig. 10—Asymmetry of control between two adjacent elements.

order of occurrence) of the P_1 and P_2 modulation is reversed, the direction of control is reversed.

To prevent signals other than those from next neighbors from forcing synchronization with the wrong phase an attenuator is placed in the signal path between each element. Stray signals will have had to pass through at least two attenuators, instead of only one, so that unwanted signals will not have a major influence in determining the phase of the subharmonic response. The natural margins of the process enable this to work properly.

With the simple combination of elements described above, not much can be done. If, however, three classes of elements are used as shown in Fig. 11, and the timing principles discussed above are employed, all logical operations can be realized. Classes of elements are defined by the timing of the modulation of the ac power supply, as shown in the figure.

The elements of classes A, B and C are shown interconnected by signal paths. Their respective power supplies have modulation as illustrated. The delay between elements is assumed to be negligible with respect to the power-supply modulation period. *Recalling that the phase of the response is determined by the phase of signals from other elements present when the power supply passes the critical level (the heavy dots in the diagram), observe at each dot which other element is* ON. For example at 1 in the diagram P_a is passing the critical level, element B is off, and element C is on. Therefore, C controls A.

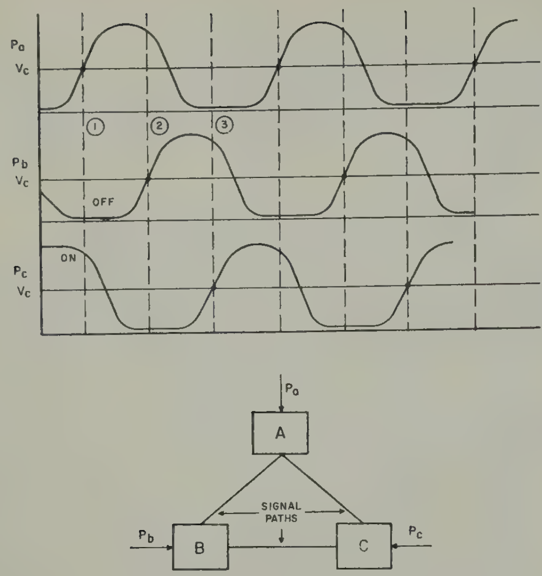


Fig. 11—Three elements classes as determined by the power supply modulation of each.

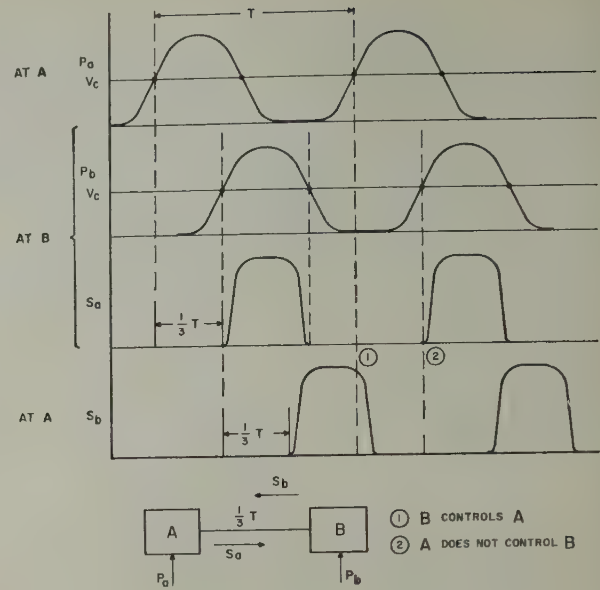


Fig. 12—Inversion of order of control.

Similarly, at ②, A controls B, and at ③, B controls C. Putting a delay between elements equal to one-third of the period of the power-supply modulation, as in Fig. 12, inverts the order of control. The exact fraction of a cycle needed to do this is related to the number of different types of elements (*i.e.*, A, B, C, $\cdots$). Three is the minimum number of types required, and results in the simplest hierarchy of control. This order of control is independent of which phase is induced in the controlled element by the controlling element. Similarly the number of classes of elements is independent of the number of possible phase states.

The delays referred to above are in terms of *group velocity* with respect to the power-supply modulation period. This period has been assumed to be long compared to the period of the sinusoidal signals involved, so that adjustments of delay to fractional cycles of the sinusoidal voltage in the signal channel will not affect appreciably the delays previously discussed. Consider now the *phase velocity* of the signal channel. The phase induced in the controlled element will depend on the phase delay between it and its controlling element. A phase Φ_i in the controlling element after a phase delay K would put the controlled element into phase state Φ_i+K. For the binary case, the phase shift K would be either an even or an odd multiple of π.

Up to this point all discussion has admitted the possibility of n phase states, or equivalently that $f_1=nf_0$, where n is an integer. The elements can thus be used to implement n-state logic. At this point complete generality will be dropped, and we shall proceed to illustrate how these principles can be applied to built up binary logic with $n=2$. With minor changes in the illustrations an n-state logic can also be realized.

Collecting nomenclature and graphical conventions which have been used, a lexicon of terminology for $n=2$ is shown in Fig. 13.

Majority Logic

The chief value of the aggregates of subharmonic generators is that they can be used to perform majority logic. By definition, a majority organ is a device or circuit which has multiple inputs and a single output. The value of the output is the value of the majority of the inputs. To avoid the indeterminate case, there must be an odd number of inputs.

In Fig. 14 the linear addition of signals from three A elements is applied to a B element.

Let

$$S_{a1} = E\cos(\omega_0 t + \Phi_1)$$
$$S_{a2} = E\cos(\omega_0 t + \Phi_2)$$
$$S_{a3} = E\cos(\omega_0 t + \Phi_3)$$

where $\Phi_1, \Phi_2, \Phi_3 = 0, \pi$, depending on the state of each A element, and $\omega_0 = 2\pi f_0$, the angular frequency of the subharmonic response. Possible results for S_a are:

$E\cos(\omega_0 t + 0)$	two $\Phi_i = 0$, one $\Phi_i = \pi$
$E\cos(\omega_0 t + \pi)$	two $\Phi_i = \pi$, one $\Phi_i = 0$
$3E\cos(\omega_0 t + 0)$	$\Phi_1 = \Phi_2 = \Phi_3 = 0$
$3E\cos(\omega_0 t + \pi)$	$\Phi_1 = \Phi_2 = \Phi_3 = \pi$.

Since the information is carried in the phase, S_a' has either 0 or π phase depending on the majority of the phases of S_{a1}, S_{a2}, and S_{a3}, and since the phase of S_b is determined by the phase of S_a' then $B=\text{Maj}(A_1, A_2, A_3)$. The amplitude variation has no importance. The truth table and equivalent logical expression are as in Fig. 15, arbitrarily letting $\Phi=0$ be state "0" and $\Phi=\pi$ be state "1."

The majority organ and a negation operation (as shown in Fig. 13) are sufficient to build all logic.

The only thing lacking at this point is a method for putting information into such a system. Permanent

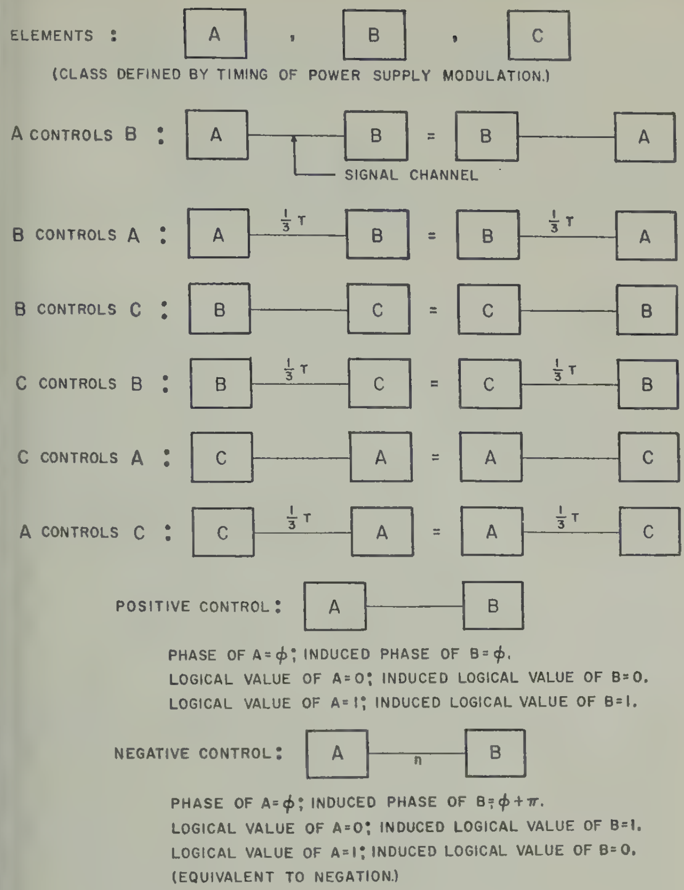


Fig. 13—Nomenclature and graphical conventions.

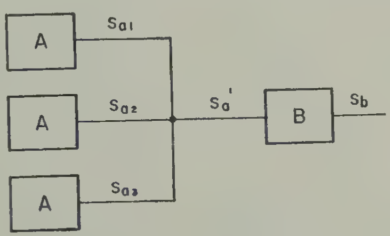


Fig. 14—Majority organ.

A_1	A_2	A_3	B
0	0	0	0
0	0	1	0
0	1	0	0
0	1	1	1
1	0	0	0
1	0	1	1
1	1	0	1
1	1	1	1

$B = A_1 \cdot A_2 \cdot A_3 + (A_1 \cdot A_2 + A_1 \cdot A_3 + A_1 \cdot A_3)$

WHERE "•" IS INTERSECTION,

AND "+" IS CONJUNCTION.

Fig. 15—Truth table for majority organ.

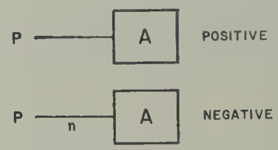


Fig. 16—Permanent sources.

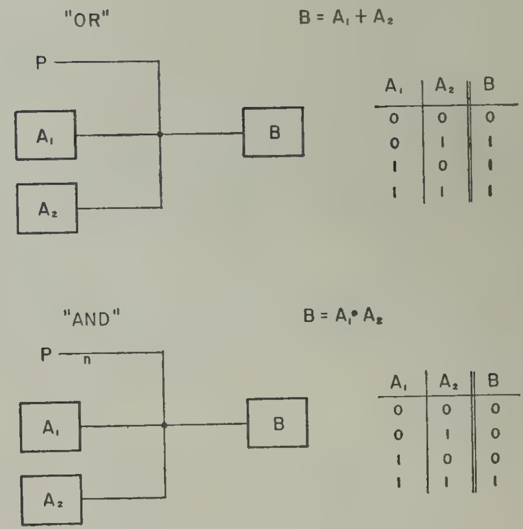


Fig. 17—Elementary logic operations.

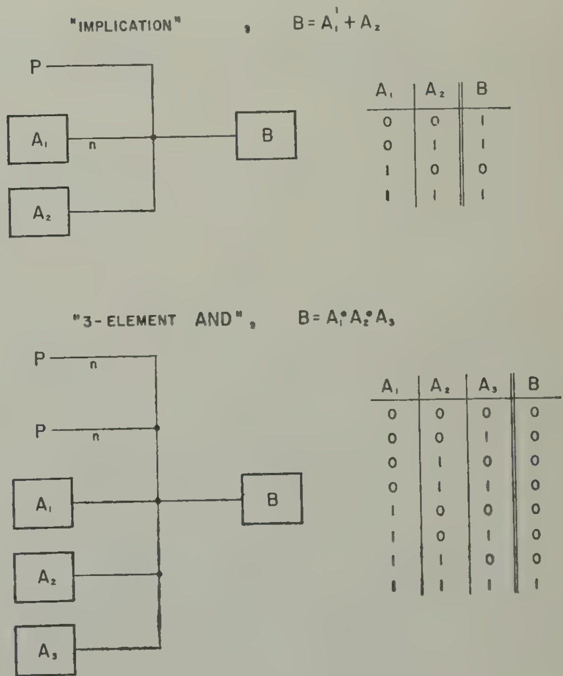


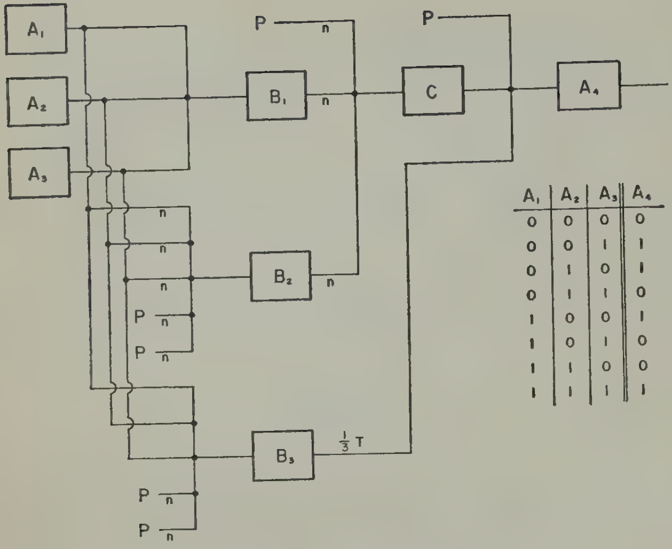
Fig. 18—Examples of nonelementary logic operations.

sources at the f_0 frequency which are gated on when desired by external controls, fill this need. A permanent source can control any class (A, B, C) of element. There are two possible types of permanent sources as illustrated in Fig. 16. The reference permanent source p is assumed to have state "1."

By using majority organs, negation, and permanent sources, the elementary logical operations shown in Fig. 17 can be performed. (At this point the separate designation of S_a as being a response of an element, A, will be dropped. The logical state represented by S_a will be called simply the *state* or *binary value* of A.)

The use of the majority organ also allows realization of other nonelementary logical functions. One of its great values is that a single organ can be used to realize many logical operations by what might be called logical biasing. This is shown in Fig. 18.

Fig. 18 may be generalized in an obvious way to n



A_1	A_2	A_3	A_4
0	0	0	0
0	0	1	1
0	1	0	1
0	1	1	0
1	0	0	1
1	0	1	0
1	1	0	0
1	1	1	1

Fig. 19—Parity circuit.

elements. Also, an n-element "or" circuit differs from the "and" circuit only in having a positive rather than a negative relationship for the permanent sources. One other important aggregate is the "parity" circuit (Fig. 19).

$$B_1 = \text{Maj}(A_1, A_2, A_3) \qquad C = B_1' \cdot B_2'$$

$$B_2 = A_1' \cdot A_2' \cdot A_3' \qquad A_4 = B_3 + C$$

$$B_3 = A_1 \cdot A_2 \cdot A_3$$

$$A_4 = A_1 \cdot A_2 \cdot A_3 + \{[\text{Maj}(A_1, A_2, A_3)]' \cdot (A_1' \cdot A_2' \cdot A_3')'\}$$

$$= A_1 \cdot A_2 \cdot A_3 + [(A_1 \cdot A_2 \cdot A_3) \cdot (A_1' \cdot A_2' \cdot A_3') + (A_1 \cdot A_2) + (A_2 \cdot A_3) + (A_1 \cdot A_3)]'.$$

As an example of the performance of a specific non-elementary logical operation, we show the majority organ and the parity circuit combined to make one stage of an adder (Fig. 20).

Reviewing what has been discussed: after establishing that the phase of the subharmonic response of a "tuned" nonlinear reactance can be used to represent logical states, the processes of negation and majority were shown to arise naturally. From these two operations the basic logical elements "and" and "or" were shown to be possible, along with other more complex logical functions from which can be built all logical operation. The entire discussion was in terms of binary computation, although it can be generalized to n-valued logic. An added attraction of majority logic is that the function of an aggregate of elements can be changed by processes implicit in the programming. For example, an "or" circuit becomes an "and" circuit by shifting the phase of the permanent source or, equally, by biasing the specific majority circuit with the output of another element involved in the computation. This provides great flexibility.

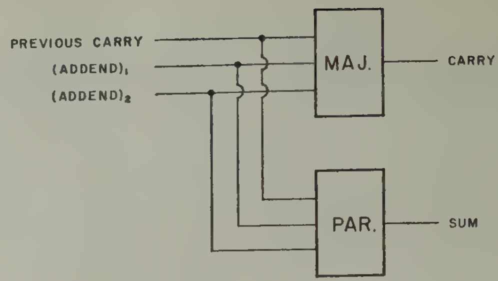


Fig. 20—One stage of an adder.

Some Practical Considerations

A typical sequence of events in a signal channel is illustrated in Fig. 21. This shows a peculiar type of AM-phase-modulated sine wave. Information is represented by phase states Φ_1 and Φ_2, which may be the same or different. The AM has no purpose in the logical processing, and so may be used to monitor the operation of the machine and, in servo-control circuits, to maintain signal levels at the proper average. The phase reference of the entire machine is the phase of the master oscillator supplying or controlling the power-supply signals for each element. The modulation of the power supply determines the basic logical time-cycle.

The modulation envelope of the power supply has been assumed to have time variations that are slow compared to the periods of both f_0 and f_1. To give a numerical example, let the period of the power supply modulation be 50 cycles of f_1 (25 cycles of f_0) in a binary circuit. Reasonable values of f_1 and f_0 are 20 kmc and 10 kmc, respectively. The basic computing cycle has, therefore, a period of 2.5 mμsec or a clock rate of 400 mc. The envelope of the modulation of both frequencies is not sinusoidal, but good rectangular pulses are not required. Three harmonics would be sufficient. Thus the modulation envelope would involve frequencies of 400–1200 mc.

To achieve a computing period of 1 mμsec (a "clock rate" of 1000 mc) with the above time ratios would require $f_0 = 25$ kmc and $f_1 = 50$ kmc. The modulation envelopes would contain frequency components between 1000 and 3000 mc. For an absolute bandwidth of 2000 mc, this would be 4 per cent and 8 per cent relative bandwidth at f_1 and f_0, respectively.

Hardware realization of this scheme may well seem outrageous to computer-systems engineers who now are in the transistor age. And it would truly be outrageous in size, cost, and power, with conventional waveguide components at X-band (8.6–12.4 kmc) and below. The techniques worked out with these conventional, readily available components can be applied to more practical geometries at a wide range of microwave frequencies. The higher frequencies of course have bad characteristics, such as high transmission attenuation, extreme precision requirements, and the lack, at present, of a complete line of components. These are things that must be overcome to push the speed of computation to the ultimate.

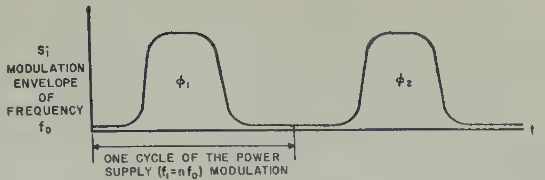


Fig. 21—Representation of PM-AM wave in the signal channel.

Von Neumann who played the leading role in the birth of the modern electronic computing machine, has, in these ideas, made another great contribution to the field. Whether this contribution will have as much importance as his original efforts can be decided only after the technology for implementing the subharmonic response scheme has been more fully worked out. At this point the prospects look good.

The above estimates of the number of cycles of the power supply and signal oscillations required for each computing cycle were made by von Neumann. Experiments using analog computer simulation of the subharmonic oscillators and LF (1–10 mc) lumped circuit models have shown that these estimates are realistic although they may be somewhat conservative. Circuits using that number of oscillation cycles per computation cycle would certainly work, and a reduction by a factor of 2.5 to 10 cycles of the signal and 20 cycles of the power supply appears to be possible. The chance of reducing the required number of cycles very much beyond this is doubtful.

Note: The Japanese have developed a subharmonic oscillator computer based on nonlinear inductance which uses the same phase script for information representation and majority logic schemes as represented in the von Neumann ideas described in this paper.[6–13] The basic circuit, called the Parametron, uses RF frequencies of 1 and 2 mc and has a computing rate of 10 kc. These two efforts are almost identical in concept, although far different in the speed of the suggested implementation, and, from available records, they seem to have been proposed in the same year, namely 1954. However, there is no direct connection between them known to this author.

[6] S. Muroga, "Elementary principle of Parametron and its application to digital computers," *Datamation*, vol. 4, No. 5, pp. 31–34; September/October, 1958.
[7] E. Goto, "On the application of parametrically excited nonlinear resonators," *Denki Tsushin Gakkai-shi;* October, 1955.
[8] E. Goto, "New Parametron circuit element using nonlinear reactance," *KDD Kenkyu Shiryo;* November, 1954.
[9] S. Oshima, "Introduction to Parametron," *Denshi Kogyo*, vol. 4, No. 11, p. 4; December, 1955.
[10] S. Oshima, "General remarks on a Parametron circuit," *Denshi Kogyo*, special volume.
[11] H. Takahashi, "The Parametron," *Tsugakkat Shi*, vol. 39, No. 6, p. 56; June, 1956.
[12] H. Yamada, "A Parametron circuit examined from the point of mathematical logic," *Denshi Kogyo*, special volume.
[13] Ohima, Enemoto, and Watanabe, "Oscillation theory of Parametron and method of measuring nonlinear elements," *KDD Kenkyu Shiryo;* November, 1955.

Stored Charge Method of Transistor Base Transit Analysis*

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***Summary*—A base layer transit time analysis has been made for high-frequency transistor base donor distributions. Transit time is defined as stored charge per unit emitter current. The emphasis on the stored charge/current ratio is particularly pertinent to high-frequency performance and facilitates qualitative analyses. The analysis applies to a *p-n-p* transistor in which the base donor density at the emitter (which specifies emitter breakdown voltage and emitter capacity for an alloyed emitter) and total number of donors per unit area of the base (which determines base resistance and emitter to collector punch-through voltage) are specified. It is shown that shorter transit times result with retarding fields since smaller base thicknesses are possible. It is thus shown that a built-in field is of lesser importance in determining transit time than is base thickness.**

* Original manuscript received by the IRE, October 29, 1958.
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Introduction

THE USE of diffusion techniques in transistor fabrication[1] has resulted in transistor designs with cut-off frequencies in the range of 1000 mc.[2] In such transistors, basic design limitations[3] seriously intrude upon freedom of design. For this reason it becomes increasingly important that each limitation be

[1] C. A. Lee, "A high frequency diffused base germanium transistor," *Bell Sys. Tech. J.*, vol. 35, pp. 23–34; January, 1956.
[2] C. H. Knowles and E. A. Temple, "Diffused base transistors," *Electronic Design*, vol. 6, pp. 12–15; July 9, 1958.
[3] J. M. Early, "Design theory of junction transistors," *Bell Sys. Tech. J.*, vol. 32, pp. 1271–1312; November, 1953. Also "Structure-determined gain-band product of junction triode transistors," Proc. IRE, vol. 110, pp. 1924–1927; December, 1958.

examined carefully in the approach to optimum designs. This paper considers the factors affecting base transit time τ of injected carriers in a diffused base junction transistor with an alloyed emitter. For convenience, *p-n-p* transistors are considered.

In the analysis presented in this paper, base transit time is defined as the ratio of stored or injected base charge to emitter current. In this sense it is more properly described as the base transit time constant. This stored charge/current ratio is important because it is directly related to high-frequency transistor performance. Recognition of the identity between stored charge and transit time allows a simple physical understanding of the qualitative features of base impurity distributions.

The base charge control concept of transistor operation is a particularly powerful one and is not generally appreciated. The first explicit exposition of this method in the literature was given by Beaufoy and Sparkes,[4] who point out that the junction transistor is fundamentally a base charge-controlled device rather than a current controlled device. While the dc characteristics can be described in terms of current control, the ac or transient characteristics are determined by the requirement for changing the charge distribution.

Reductions in base transit time[5,6] in a transistor with a given base width w can be obtained with aiding electric fields resulting from the grading of base impurity densities. This principle has been widely used[1,2,7] with the advent of diffusion techniques. The limitations imposed in the highest frequency designs employing alloyed emitters do not permit advantage to be taken of this phenomenon, and shorter transit times are possible with retarding fields. This seemingly paradoxical result is made understandable by demonstrating that requirements for adequate emitter breakdown and emitter to collector punch-through voltages, and for minimum emitter junction capacity and base resistance, lead to narrower base widths and lower transit times if retarding fields are accepted.

Base Transit Time

The concept of base transit time is useful in analyzing high-frequency performance of transistors. While it sheds no light directly on the broadening of an input signal as do the ac solutions of carrier flow of Kromer,[5] it is particularly valuable in approximate and comparative analyses because of its conceptual simplicity.

On closer examination this simplicity has not always been apparent. Signal transmission cannot be thought of rigorously as a time of flight phenomenon. This is inherent in the diffusion mechanism and mathematically is a consequence of the diffusion equation which possesses no traveling-wave solutions. Moll and Ross[6] calculated transit time by defining a velocity v associated with the carrier flow through the relation

$$J = qpv$$

where J is current density, q electronic charge, and p hole density. While this yields a correct result, the procedure is essentially mathematical because v is not a real velocity having a direct physical interpretation.

The conceptual difficulties are eliminated, however, by defining transit time as the ratio of stored or injected charge Q_s to emitter current I_e

$$\tau = \frac{Q_s}{I_e}$$

or more generally

$$\tau = q \int_0^w \frac{p dx}{J}, \tag{1}$$

where w is base width and J is emitter current density. It will be noted that the ratio of stored charge to current is the average time spent per carrier in the base and this is the principal reason for defining it as transit time. While this procedure is formally equivalent to that of Moll and Ross, the point of view differs because it focuses attention on the stored-charge/current ratio as the important physical concept.

The importance of this point of view can be demonstrated by showing that the ratio of high-frequency base current I_{b1} to emitter current I_{e1} is directly proportional to the stored charge/current ratio. In a transistor having unity dc α, the base stored charge can change only through base current flow. This is related to Q_{s1}, the ac component of stored charge, $I_{b1} = j\omega Q_{s1}$, giving

$$\frac{I_{b1}}{I_{e1}} = j\omega \frac{Q_{s1}}{I_{e1}}$$
$$= j\omega\tau.$$

This current ratio, which is proportional to the charge/current ratio or transit time constant τ, should be minimized for the most favorable high-frequency performance. We have used the familiar low-frequency approximation ($\omega < 1/\tau$) of a modulated dc hole distribution for which the ratios of ac quantities are equal to the ratios of the corresponding dc quantities.

As demonstrated in the analyses to follow, consideration of stored charge can often simplify problems by substituting physical reasoning and intuition for detailed calculations. The effects of complicated variations of donor distributions on transit time are analyzed in this way.

[4] R. Beaufoy and J. J. Sparkes, "The junction transistor as a charge-controlled device," *ATE J.*, vol. 13, no. 4, pp. 310–327; 1957.

[5] H. Kromer, "Zur theorie des diffusions—und des drift transistors," *Arch. elek. Übertragung*, vol. 8, pp. 223–228, 363–369, 499–504; May, August, November, 1954.

[6] J. L. Moll and I. M. Ross, "The dependence of transistor parameters on the distribution of base layer resistivity," Proc. IRE, vol. 44, pp. 72–78; January, 1956.

[7] H. Kromer, "Der Drifttransistor," *Naturwiss.*, vol. 40, pp. 578–579; November, 1953.

Solutions for Carrier Flow

Eq. (1) requires knowledge of base hole distribution for calculation of transit time. The general solutions for carrier transport through a base region from $x=0$ at the emitter to $x=w$ at the edge of the collector depletion layer have been given by Moll and Ross.[6] Eqs. (2) through (5) have been taken from their paper.

The hole current density is given by

$$J = q\mu_p pE - qD_p \frac{dp}{dx}, \qquad (2)$$

where μ_p and D_p are mobility and diffusion coefficients for holes in the base and p is hole density. Recombination in the base has been neglected. The built-in field is given by

$$E = -\frac{kT}{q}(1/N)dN/dx, \qquad (3)$$

where N is donor density. Solution of (1) and (2) gives the hole density

$$p = (J/qD_p)(1/N(x))\int_x^w Ndx. \qquad (4)$$

The lower limit $x=0$ defines the hole density at the emitter p_0

$$p_0 = (J/qD_pN_0)\int_0^w Ndx \qquad (5)$$
$$= (J/qD_pN_0)N_T,$$

where N_0 is the donor density at the emitter ($x=0$) and N_T, given by

$$N_T = \int_0^w Ndx, \qquad (6)$$

is the total number of donors per unit area of the base.

Conditions Imposed on High-Frequency Transit Time Analysis

Freedom to choose base layer distribution of impurities is more severely limited in high-frequency designs than in low-frequency designs. In addition to the usual requirement for a high emitter doping relative to base doping, several specifically high-frequency conditions are encountered. For convenience, a p-n-p transistor will be considered.

If the base doping level does not change appreciably in the emitter space charge region, the usual case, the emitter breakdown voltage[8] and the junction capacity C_{TE},[9] are specified by the base doping level at the emitter. Since the current which flows in C_{TE} is not part of of the injected base current, C_{TE} must be limited in order to have adequate high-frequency current gain. These considerations lead to a specification in the subsequent analysis of a maximum doping level at the emitter N_0.

The total number of impurities per cm² of the base, N_T, introduced in the preceding section, specifies both the transverse base resistance and the punch-through voltage. The base resistance directly affects the high-frequency figure of merit.[3] This requirement as well as the necessity for an adequate punch-through voltage lead to a specified value N_T for the total number of impurities per cm² in the base.

With the specification of N_0 and N_T the problem then is how to distribute the donors to minimize base transit time. How this is done will determine base thickness w. Note from (5) that the hole density at the emitter p_0 is expressed in terms of J, N_0, and N_T. Thus if J is assigned a fixed value, any two of the three parameters, p_0, N_0, and N_T can be selected to have specified values.

Qualitative High-Frequency Transit Time Analysis

In this section a qualitative analysis is made of the effect of different types of donor distributions on transit time. It relies solely on a consideration of stored charge.

Three donor distributions for which N_0 and N_T are the same are shown in Fig. 1(a). The uniform distribution has width w_0 while the two simple exponential distributions giving rise to a retarding and to an aiding field have, as required, smaller and larger values of w, respectively. The corresponding solutions to p of Fig. 1(b) for equal values of emitter current J can be sketched without recourse to calculation. The solution for the uniform distribution is known to be linear. Each distribution has the same value p_0 at $x=0$ for equal current density J as noted in the previous section. Equal values for J require equal slope for each curve at $x=w$ ($p=0$) from (2). With this information the curves of Fig. 1(b) follow.

The area under each curve of Fig. 1(b) is the stored charge and is proportional to transit time. Thus, the narrower retarding field distribution gives the smallest value of transit time, and the wider aiding field distribution gives rise to the greatest.

This qualitative procedure can be applied to more complicated cases. The retarding field case (A) just discussed will be compared in Fig. 2(a) with a distribution (B) having simultaneously both aiding and retarding fields while still having equal values for N_0 and N_T. The B distribution is more nearly a physically realizable distribution which might result from the out-diffusion of an initially high concentration very close to the surface. The discontinuity at $x=w$ corresponds to the edge of the collector space charge depletion layer. At $x=x_m$, the maximum of the donor curve, there is no built-in

[8] S. L. Miller, "Avalanche breakdown in germanium," *Phys. Rev.*, vol. 99, pp. 1234–1241; August 15, 1955.

[9] W. Shockley, "The theory of p-n junctions in semiconductors and p-n junction transistors," *Bell Sys. Tech. J.*, vol. 28, pp. 435–489; July, 1949.

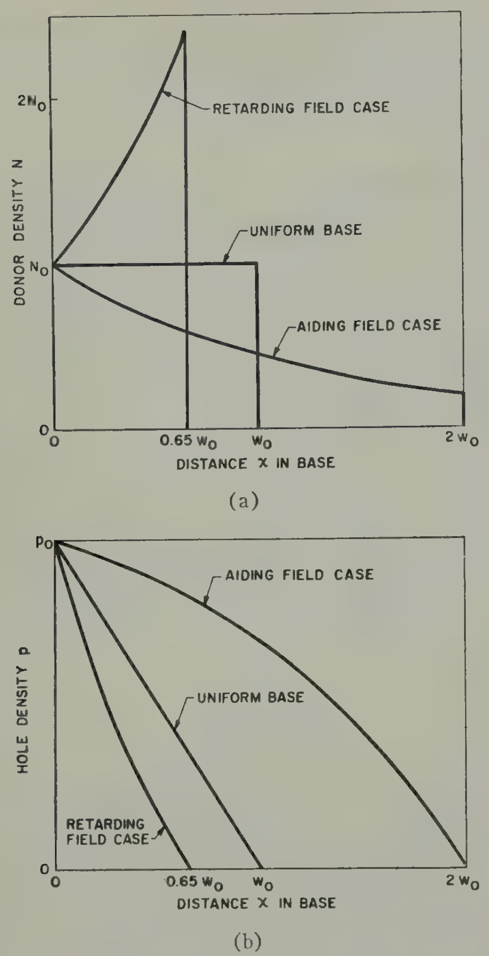


Fig. 1—(a) Comparison of donor distributions for uniform base, retarding and aiding field cases. (b) Comparison of hole distributions for uniform base, retarding and aiding field cases.

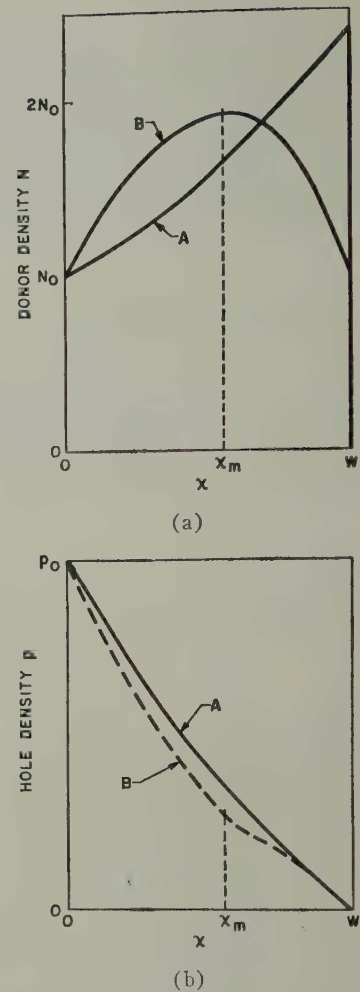


Fig. 2—(a) Comparison of A and B donor distributions. (b) Comparison of hole distributions for A and B donor distributions.

field while a retarding field exists for $x<x_m$ and an aiding field for $x>x_m$.

The qualitative features of the density curve for holes p can be readily seen for each of these distributions. The curve for the simple retarding exponential A distribution is taken from Fig. 1(b). For the same reason as in the comparisons made in Fig. 1 and as discussed above, the slopes of the curves at $x=w$ and the values for p_0 must be identical for the A and B distributions. In addition, the slope for the B distribution at $x=x_m$ (the zero field point) must be the same as at $x=w$. The slope in the range $x_m<x<w$ (for which $E>0$) must be less than at $x=w$, as can be seen from the aiding field case discussed in connection with Fig. 1(b). The hole curve for this B distribution is now readily constructed, starting at $x=w$. The slope here is identical with the A curve, but decreases initially as x approaches x_m because of the aiding field. The slope then increases again so that the slope at $x=x_m$ is the same as at $x=w$. For $x<x_m$ the slope increases smoothly as the curve approaches p_0 at $x=0$. Thus it is seen that the area under the hole density curve or stored charge corresponding to the B distribution and consequently the transit time is less than for the A distribution.

Quantitative Analysis for Exponential Distributions

The qualitative results for the exponential distribution of donors will be demonstrated analytically. The base donor distribution is given by

$$N = N_0 \exp(-\beta x). \tag{7}$$

From (2) the field is given by $E=\beta(kT/q)$, showing that $\beta>0$ and $\beta<0$ correspond to aiding and retarding fields, respectively. Distributions of this type were given in Fig. 1(a).

The condition for a specified value of N_T (6) leads to a relation between w and β,

$$\frac{w}{w_0} = \frac{\ln(1/(1-\beta w_0))}{\beta w_0}, \tag{8}$$

where w_0 is the width corresponding to uniform doping.

From (4) the hole density is given by

$$p = \frac{J}{qD_p\beta}\left[1 - \exp(\beta(x-w))\right]. \tag{9}$$

From (1) the transit time is given by

$$\tau = \frac{w_0^2}{2D_p} f, \tag{10}$$

which is the product of the transit time for a uniform base times a base transit time factor f given by

$$f = 2\frac{w/w_0 - 1}{\beta w_0}. \tag{11}$$

It will be recalled that this solution applies, as described earlier, to donor distributions with specified values of N_0 (donor density at the emitter) and N_T (total number of donors per unit area).

Because of the transcendental character of (8), it is not possible to express βw_0, and hence τ and f, as a function of w/w_0. However, this function is presented graphically in Fig. 3. As expected, f is unity for $w/w_0 = 1.0$ ($\tau = w_0^2/2D_p$ for a uniform base).

It is interesting to note that the base transit time factor f is given approximately by

$$f \cong (w/w_0)^{4/3} \tag{12}$$

to within a few per cent for $0.1 \leq w/w_0 \leq 6$. Expansion of f in powers of $(w/w_0) - 1$ shows that the first and second derivatives of f are identical with those for the 4/3 power law and the third derivative is very nearly the same.[10] The effect of built-in fields can be appreciated by comparing this with

$$f = (w/w_0)^2$$

for a uniform base with no built-in fields and no restriction on N_T.

So far, variation of diffusion coefficient with doping level has not been taken into account. For $N \geq 10^{16}$, the diffusion coefficient[11] in germanium varies roughly as $N^{-1/3}$. In a crude way the average donor density varies as w_0/w, so that D_p varies as

$$D_p \cong D_{p0}(w/w_0)^{1/3}$$

where D_{p0} is the diffusion coefficient applicable to the uniformly doped base. Thus the base transit time is given roughly by

$$\tau \cong \frac{w_0^2}{2D_{p0}} (w/w_0). \tag{13}$$

Eqs. (12) and (13) show that considerable advantage in reducing transit time can be obtained by making base thickness less than w_0 even though retarding fields result.

Conclusion

An analysis has been made of the effects on transit time required by high-frequency designs. This has been done by defining base transit time as the ratio of stored base charge to emitter current. In this way it can be interpreted as the average time spent per carrier in the base. Simple physical reasoning applied to this stored charge point of view provides considerable insight into the factors affecting transit time.

This method of analysis has been applied to the design of base layer impurity distributions of high-frequency transistors. By requiring that the base donor density at the emitter (which determines breakdown voltage and emitter capacity with alloyed emitters) and the total number of donors per unit area of the base (which determines base resistance and emitter to collector punch-through voltage) be specified, it is shown that distributions having retarding fields yield shorter transit times than do those with aiding fields. This result stems from the narrower base thickness possible with retarding fields for these boundary conditions. Thus it can be concluded that the effects of built-in fields are minor compared to the actual value of base thickness and that in high-frequency designs the actual value should be reduced even though retarding fields result.

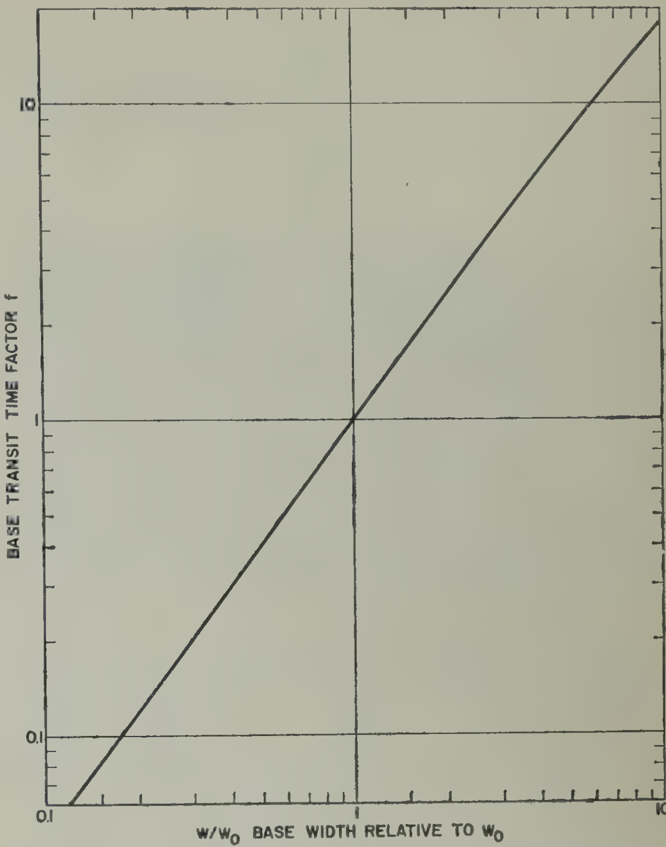


Fig. 3—Base transit time factor as a function of base thickness.

Acknowledgment

The author wishes to thank J. M. Early for many stimulating discussions and helpful suggestions during the course of this work. He also wishes to thank I. M. Ross for helpful comments on the manuscript.

[10] The author is indebted to F. W. Terman for analyzing the equations.

[11] M. B. Prince, "Drift mobilities in semiconductors. I. Germanium," *Phys. Rev.*, vol. 92, pp. 681–687; November 1, 1953.

The Hall Effect Circulator—a Passive Transmission Device*

W. J. GRUBBS†

Summary—**Three-port nonreciprocal Hall effect devices have been made which circulate dc and ac signals either in a clockwise or counterclockwise sense. Forward losses of 17 db and reverse losses of 61 db have been obtained, giving a transmission ratio of 44 db. With the aid of simple six-resistor networks (or even simpler three-resistor networks) the nine short circuit admittance parameters of the circulator can be adjusted in a calculable manner. These networks permit asymmetrical circulators to appear symmetrical, to operate over a wide range of impedance levels, to operate with any value of magnetic field, and to introduce no loss—or even gain—if negative resistances are used. An analysis of the circulator, with and without the parallel networks, is included. It is shown that the minimum possible forward loss for a Hall effect circulator is 8.4 db.**

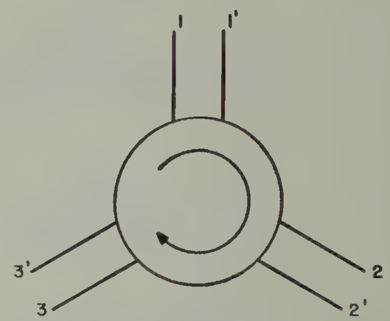


Fig. 1—Symbol for three-port circulator.

Introduction

Most of the passive transmission devices one normally encounters are reciprocal in nature; that is, they exhibit the same transfer characteristic whether a signal is applied at some point A and is removed at point B, or vice versa. A list of such devices might include transformers, filters, equalizers, transmission lines, etc. Some passive transmission devices, however, are not reciprocal. In these devices, the nature of the transfer characteristic depends on whether the signal goes through from A to B or from B to A. Gyrators, isolators, and circulators are in the category of nonreciprocal devices and have been made for use in waveguides for several years.

In 1953, it was reported that Hall effect gyrators and isolators could be made for use in wire circuits.[1] These latter devices function throughout the frequency range from dc to some upper limit which depends theoretically only on the dielectric relaxation time of the material used. The purpose of this paper is to describe and analyze the Hall effect circulator, which has a similar frequency range. This device was originally conceived by Semmelman[2] of Bell Laboratories.

The Hall Effect Circulator

A circulator is a three-port (six-terminal) nonreciprocal passive device which circulates signals essentially in one direction only—clockwise, for example, as shown in Fig. 1. Thus, an input signal on terminals 1-1′ is transmitted to terminals 2-2′, but no signal appears at 3-3′. The same clockwise circulation occurs when a signal is applied to any of the three terminal pairs.

If a magnetic field is applied perpendicular to a current flow (see Fig. 2), the current carriers are deflected sidewise and a transverse electric field (perpendicular to both the magnetic field and the longitudinal current flow) is built up between the sides of the conductor. Thus the net electric field in the sample makes some angle θ with the direction of current flow. This angle is known as the Hall angle and, to a first approximation, its tangent is given by the product of the magnetic field and the Hall mobility of the majority carriers in the sample when the product of the majority carrier density and mobility is much greater than the same product for minority carriers. Hall effect devices use semiconductors rather than metals because most metals have much lower mobilities.

If six equally-spaced contacts are made to the edge of a slice of semiconductor, as shown in Fig. 3, and a constant magnetic field is applied perpendicular to the plane of the slice, the slice becomes a Hall effect circulator for the proper value of magnetic field. That is, θ must be such that the net electric field produces no voltage between terminals 3 and 3′.

As soon as a load is attached to terminals 2-2′ so that a current can flow, an additional Hall field (not shown here) is produced which has the effect of decreasing the effective Hall angle in the sample. Consequently, a stronger magnetic field is required to balance the circulator. The value of the magnetic field required for a matching load impedance will be called B_0. Thus, the circulator can be made to operate properly with a range of load resistances by adjusting the magnetic flux density.

Analysis of Circulator

Most n-port circuits or devices can be completely described and characterized by n^2 parameters. These parameters may be either open-circuit impedances or short-

* Original manuscript received by the IRE, October 29, 1958; revised manuscript received, December 22, 1958.

† Bell Telephone Laboratories, Inc., Murray Hill, N. J.

[1] W. P. Mason, W. H. Hewitt, and R. F. Wick, "Hall effect modulators and 'gyrators' employing magnetic field independent orientations in germanium," *J. Appl. Phys.*, vol. 24, pp. 166–175; February, 1953.

[2] C. L. Semmelman, "Nonreciprocal Transmitting Devices," U. S. Pat. 2,774,890; December 18, 1956.

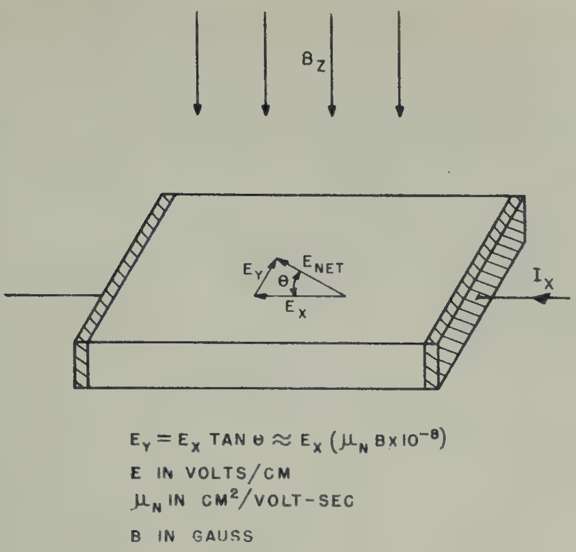


Fig. 2—Illustration of the Hall effect.

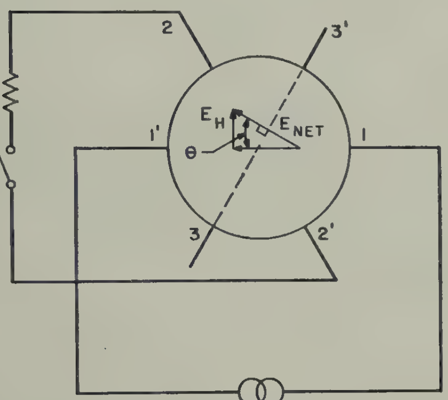


Fig. 3—Diagram showing qualitatively how the Hall field (E_H) causes 3 and 3′ to be on the same equipotential line with a signal applied between 1 and 1′.

circuit admittances. Although impedances are more easily measured, in the circulator it is preferable to work with the short-circuit admittances. In practice, admittances are computed from measured impedances.

These admittances are defined as:

$$y_{mn} = \frac{i_m}{v_n}\bigg|_{\text{all } v\text{'s}=0 \text{ except } v_n}.$$

The set of three equations relating v's, i's, and y's are

$$i_m = \sum_{n=1}^{3} y_{mn} v_n; \qquad m = 1, 2, 3. \tag{1}$$

One form of equivalent circuit which incorporates these admittances is shown in Fig. 4. In the case of a symmetrical circulator, the three self-admittances are equal ($y_{11} = y_{22} = y_{33} = y_S$); and three forward transfer admittances are equal ($y_{13} = y_{32} = y_{21} = y_F$); and the three reverse transfer admittances are equal ($y_{12} = y_{23} = y_{31} = y_R$).

The input admittance is given by

$$y_{IN} = y_S\left[1 + \frac{U_F^3 + U_R^3 - 2U_RU_F(1+U_L)}{(1+U_L)^2 - U_RU_F}\right], \tag{2}$$

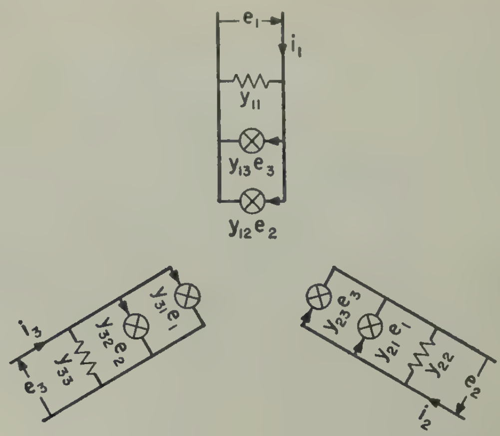


Fig. 4—Equivalent circuit of Hall effect circulator.

where the admittances have been normalized with respect to y_S, such that $U_F = y_F/y_S$, $U_R = y_R/y_S$, and $U_L = y_L/y_S$. y_L is the load admittance and is assumed to be the same at each terminal pair. These normalized admittances are useful for analysis because they will be the same for any Hall effect circulator in which θ has the same value.

The forward voltage transmission ratio (from 1-1′ to 2-2′, or from 2-2′ to 3-3′, etc.) is

$$\frac{V_F}{V_{IN}} = \frac{U_R^2 - U_F(1+U_L)}{(1+U_L)^2 - U_RU_F}. \tag{3}$$

The reverse voltage ratio (that is, for a signal transmitted in the counterclockwise direction) is

$$\frac{V_R}{V_{IN}} = \frac{U_F^2 - U_R(1+U_L)}{(1+U_L)^2 - U_RU_F}. \tag{4}$$

In a circulator V_R must be zero. Thus, the condition for circulation may be stated as

$$U_F^2 - U_R(1+U_L) = 0. \tag{5}$$

Fig. 5 shows how the admittances vary in an n-type Ge circulator with the magnetic field, B. At $B=0$, $y_R = y_F$; in other words, with no magnetic field, the device is reciprocal and cannot distinguish the clockwise from the counterclockwise direction. When B has some finite value, $y_R \neq y_F$, and the device is nonreciprocal. As B increases, the device becomes more and more nonreciprocal. At any flux density greater than about 10 kilogauss, the condition for circulation can be met by proper choice of y_L. At $B=10$ kg, y_L must be zero; *i.e.*, the load terminals must be open-circuited. As B is increased, y_L must increase to satisfy (5). At the value of flux density called B_0, about 14.5 kg for n-type Ge, $y_L = y_{IN}$. Therefore, this is the logical value of field to use, because this is the field at which the load matches the admittance seen looking back into the circulator.

Using (3) and the admittance values at $B = B_0$, we can calculate the forward loss to be 17 db. The measured forward losses ranged from 16 to 18 db on the various n-type Ge circulators which were fabricated. The for-

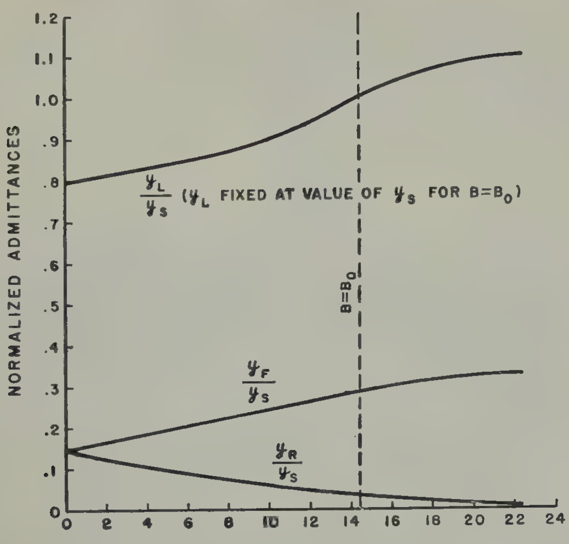


Fig. 5—Variation of short-circuit admittances with magnetic field.

ward loss will be the same for any Hall effect circulator, even for a larger or smaller sample made of a different material. The preceding statement is based on the assumptions that there are no surface effects and no asymmetrical bulk effects, in which case the Hall angle θ must always have the same value. With the same θ, the normalized forward and reverse transfer admittances will have the same value. Therefore, v_F/v_{IN} will be unchanged, assuming, of course, that U_L is given a value of one.

Because of the relatively high forward loss (17 db), reflections are severely attenuated, so that the input admittance is very nearly constant. In fact, it can be shown from (2) that $Y_{IN}=y_S\pm0.5$ per cent for any symmetrical load admittance from short circuit to open circuit.

Use of Parallel Networks

Hall effect circulators have been made which are quite symmetrical. The best circulator made to date has an average reverse loss of 61 db and an average forward loss of 17 db, giving an average transmission ratio of 44 db. Any two pairs of terminals on such a circulator would act as an isolator with 17 db loss in the forward direction. However, the symmetry depends primarily on having a regular 60° angular spacing between the six edge contacts. This is not impossible, but it is difficult, to say the least. Furthermore, it is almost impossible to obtain a specific input admittance (or operating impedance level) because, to a good approximation, this is directly proportional to the material's conductivity and the reciprocal of the sample thickness. In a null transmission device such as a circulator, impedance matching is essential.

With the aid of a resistance network connected in parallel with the circulator, imperfections can be effectively removed and several other interesting features are made available. For instance, the forward loss may be decreased by using a larger magnetic field ($B>B_0$). On the other hand, if one can tolerate greater forward loss, one can use a smaller magnet ($B<B_0$). The main requirement for circulation is that the reverse loss be theoretically infinite. This requirement can be met for the limiting case of $B=0$, but the forward loss also becomes infinite.

With any fixed B, circulation is obtainable with any load admittance from about ten times the normal admittance down to zero—*i.e.*, open-circuit. However, the admittance of the circulator with the network changes only slightly over this range, so that the forward loss becomes quite high when a widely different load admittance is used, due simply to the admittance mismatch. Furthermore, the resistance networks can provide circulation when an essentially symmetrical circulator is used with a different load conductance on each of the three terminal pairs.

If one were willing to use negative resistances in the network, one could eliminate the forward loss (which is normally about 17 db) without decreasing the maximum reverse loss in the balanced condition. In fact, one could obtain negative loss, or gain.

The six-terminal (three-port) resistance network can be represented by

$$i_m' = \sum_{n=1}^{3} y_{mn}'v_n'; \qquad m = 1, 2, 3. \tag{6}$$

If we connect the resistance network in parallel with the circulator by connecting terminals 1 and 1′ of the one to 1 and 1′ of the other, etc., then the three equations describing the combined network are

$$I_m = \sum_{n=1}^{3} Y_{mn}V_n; \qquad m = 1, 2, 3,$$

where [see (1)]

$$I_m = i_m + i_m' \tag{7}$$

$$Y_{mn} = y_{mn} + y_{mn}', \tag{8}$$

$$V_n = v_n = v_n'. \tag{9}$$

The resistance network admittances (y_{mn}') for an asymmetrical network become so bulky that they are practically useless. However, if one assumes symmetry, they are (for network C_p shown in Fig. 6).

$$y_F' = y_R' = -\frac{1}{2R}$$

$$y_S' = \frac{1}{R}\cdot \tag{10a}$$

For the network shown in Fig. 7 (C_S), the admittances are

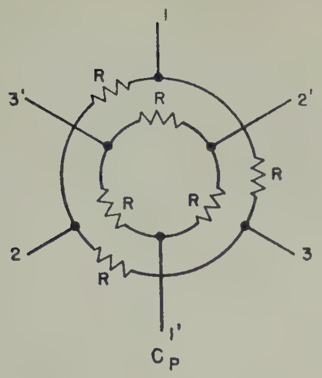


Fig. 6—Network C_p.

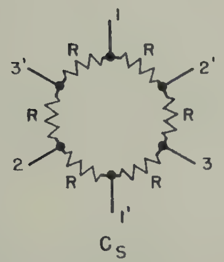


Fig. 7—Network C_S.

$$
y_F'' = y_R'' = +\frac{1}{2R}
$$

$$
y_S'' = \frac{1}{R}\cdot \tag{10b}
$$

In both cases, R is the value of the individual resistors.

Thus, by the simple addition indicated in (8), we arrive at the effective admittances of the parallel combination of circulator and resistance network C_p:

$$
Y_F{}^P = y_F + y_F' = y_F - \frac{1}{2R}
$$

$$
Y_R{}^P = y_R - \frac{1}{2R}
$$

$$
Y_S{}^P = y_S + \frac{1}{R}\cdot \tag{11}
$$

Similarly, the effective admittances when the resistance network is C_S are:

$$
Y_F{}^S = y_F + \frac{1}{2R}
$$

$$
Y_R{}^S = y_R + \frac{1}{2R}
$$

$$
Y_S{}^S = y_S + \frac{1}{R}\cdot \tag{12}
$$

The requirements for circulation (when C_p is used) may be written as

$$
(Y_F{}^P)^2 - Y_R{}^P(Y_S{}^P + y_L) = 0,
$$

or

$$
\left(y_F - \frac{1}{2R}\right)^2 - \left(y_R - \frac{1}{2R}\right)\left(y_S + y_L + \frac{1}{R}\right) = 0. \tag{13}
$$

If we define $R = 1/ky_S$, we can use normalized admittances in (13):

$$
\left(U_F - \frac{k}{2}\right)^2 - \left(U_R - \frac{k}{2}\right)(1 + U_L + k) = 0. \tag{14}
$$

Solving (14) for k yields

$$
\frac{3}{2}k = U_F + U_R - \tfrac{1}{2}(1 + U_L) + \sqrt{[-U_F - U_R + \tfrac{1}{2}(1 + U_L)]^2 - 3[U_F{}^2 - U_R(1 + U_L)]}. \tag{15}
$$

Eq. (15) is of the form

$$
\frac{3}{2}k = -a + \sqrt{a^2 - 3b}.
$$

It may be noted that, when $B = B_0$, $b = 0$, and $k = 0$. In other words, the parallel resistance network must be removed for the circulator to work properly. Examination of (15) and the curves of Fig. 5 shows that k is real, positive and finite only when $0 \leq B < B_0$.

The requirement for circulation (when C_S is used) is

$$
\left(y_F + \frac{1}{2R}\right)^2 - \left(y_R + \frac{1}{2R}\right)\left(y_S + y_L + \frac{1}{R}\right) = 0,
$$

which, using normalized admittances, is

$$
\left(U_F + \frac{k}{2}\right)^2 - \left(U_R + \frac{k}{2}\right)(1 + U_L + k) = 0. \tag{16}
$$

Solving (16) for k yields

$$
\frac{k}{2} = U_F - U_R - \tfrac{1}{2}(1 + U_L) + \sqrt{[-U_F + U_R + \tfrac{1}{2}(1 + U_L)]^2 + [U_F{}^2 - U_R(1 + U_L)]}. \tag{17}
$$

Eq. (17) is of the form

$$\frac{k}{2} = -c + \sqrt{c^2 + b}.$$

Again, when $B = B_0$, $b = 0$, and $k = 0$. Studying (17) and Fig. 5 shows that k is real, positive and finite only when $B > B_0$. For completeness, the expressions for forward and reverse voltage ratios are given below for combined networks using C_P and C_S. The symbols have the same meanings as before.

With network C_P

$$\frac{e_F}{e_{IN}} = \frac{\left(U_R - \frac{k}{2}\right)^2 - \left(U_F - \frac{k}{2}\right)(1 + U_L + k)}{(1 + U_L + k)^2 - \left(U_R - \frac{k}{2}\right)\left(U_F - \frac{k}{2}\right)} \tag{18}$$

$$\frac{e_R}{e_{IN}} = \frac{\left(U_F - \frac{k}{2}\right)^2 - \left(U_R - \frac{k}{2}\right)(1 + U_L + k)}{(1 + U_L + k)^2 - \left(U_R - \frac{k}{2}\right)\left(U_F - \frac{k}{2}\right)} \tag{19}$$

with network C_S

$$\frac{e_F}{e_{IN}} = \frac{\left(U_R + \frac{k}{2}\right)^2 - \left(U_F + \frac{k}{2}\right)(1 + U_L + k)}{(1 + U_L + k)^2 - \left(U_R + \frac{k}{2}\right)\left(U_F + \frac{k}{2}\right)} \tag{20}$$

$$\frac{e_R}{e_{IN}} = \frac{\left(U_F + \frac{k}{2}\right)^2 - \left(U_R + \frac{k}{2}\right)(1 + U_L + k)}{(1 + U_L + k)^2 - \left(U_R + \frac{k}{2}\right)\left(U_F + \frac{k}{2}\right)}. \tag{21}$$

It is possible to use simpler forms of these resistance networks. The simplified networks (which will be called C_P' and C_S') consist of only three resistors instead of six. See Figs. 8 and 9. Either of the forms of C_S' are usable. However, the one on the right has the advantage of being symmetrical. Furthermore, the symmetrical C_S' is the only one of the three which can be analyzed by the method used on the six-resistor networks, and it is the only one of the three which permits the terminal pairs to be balanced-to-ground.

A major advantage of C_P' and C_S' (aside from the obvious one that they use only half as many resistors) is that they are much easier to adjust. For any particular desired circuit condition, there is only one set of three resistance values which will operate properly; also, if the resistance values are nearly right, the isolation between any two terminal pairs is controlled almost entirely by one particular resistor. With the six-resistor networks there is an infinity of sets of six resistance values which will cause the circuit to operate in a desired fashion. In this latter case, the resistors work in pairs and each interacts strongly with one another. Thus, each of the three pairs of resistors can take on an infinite number of paired values.

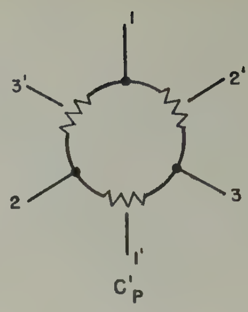


Fig. 8—Network C_p'.

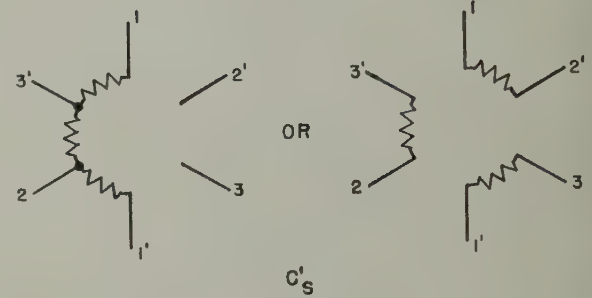


Fig. 9—Two possible forms of network C_s'.

If negative resistances are used in the resistance networks, calculations indicate that the forward loss can be reduced to zero while the infinite reverse loss is maintained.

Thus it is possible to cause a circulator to operate with any value of B field fron zero to as high a field as is attainable. This has been demonstrated with an n-type Ge circulator which, with $B = B_0$, requires a load impedance $z_L = 220\ \Omega$. So $y_L = 1/220 = 4.55$ millimhos. The measured admittances plotted in Fig. 5 were used in (15) and (17) to calculate values of k. These k values, along with the experimentally determined k's, are plotted in Fig. 10 as the k_6 curve. Considering the amount of numerical calculation involved, the agreement is quite close. The solid k_3 curve was calculated in a similar manner for the symmetrical C_S'. The expression is

$$k = \frac{3[U_F^2 - U_R(1 + U_L)]}{U_R + U_L - 2U_F + 1}. \tag{22}$$

The square points for magnetic fields of less than B_0 give measured values of k_3 in C_P'. The solid L_F curve was calculated from (18) for $B < B_0$, and (20) for $B > B_0$. The points represent experimentally determined values. B_0 is the field at which this n-type Ge circulator operates with no parallel network (14.5 kg in this instance).

(See the Appendix for a calculation of a lower bound on the forward loss.) The preceding analysis has dealt with the case of constant load admittance and varying B field. Likewise, the circulator can be made to operate with either increased or decreased load admittance and

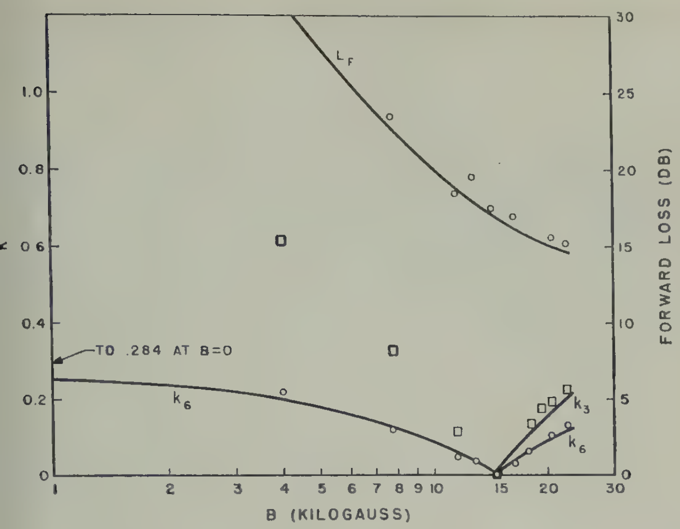


Fig. 10—Variation of L_F (forward loss) and k_3 and k_6 (normalized admittances of three-resistor and six-resistor network elements, respectively) with magnetic field; reverse loss is maintained theoretically at infinity.

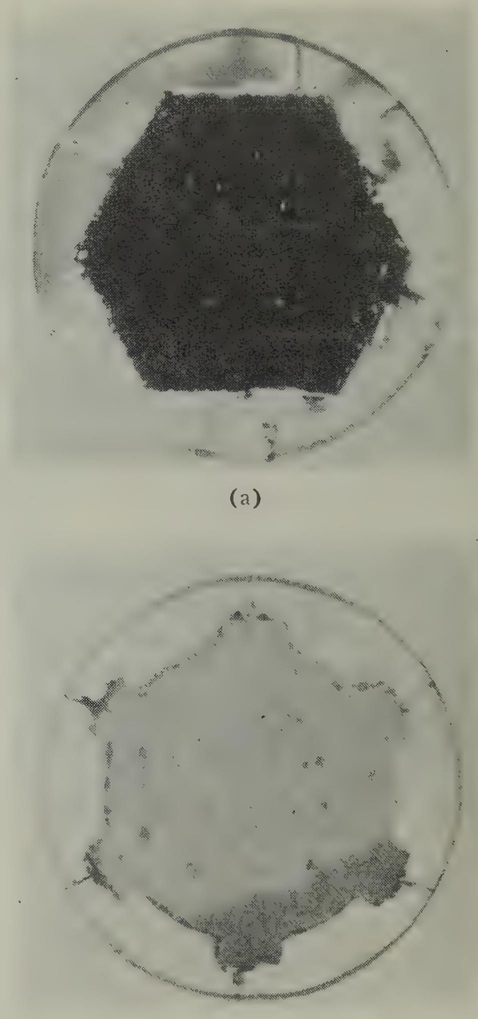
the same B field; *e.g.*, $B=B_0$. This is accomplished by varying the value of k. However, the impedance at the terminals is nearly constant in this case. Therefore, although the loads are such as to allow the circulator to function properly, there is considerable power loss due to the admittance mismatch. Nevertheless, this is a desirable method of "tuning" the admittance level of the device. If y_L is not very different from y_S, the increased forward loss will be negligible. This procedure is immensely more practical than attempting to maintain accurate control over the thickness and resistivity of the sample. Furthermore, the use of a parallel network removes the need for accurate placement of terminals.

This last sentence refers to the fact that asymmetrical circulators can be made symmetrical with the addition of a resistance network. Likewise, an essentially symmetrical circulator can be made to appear asymmetrical in some desired manner to permit the use of different load admittances at the three terminal pairs. It will be remembered that additional loss is experienced when this is done, but it might still be desirable in some instances.

Method of Construction

The degree of balance which a circulator will afford is dependent almost exclusively on the geometric placement of the six terminals. Therefore, considerable effort has been devoted to finding a "foolproof" method of achieving a regular 60° angular spacing between adjacent terminal centers.

The circulators made to date have been made from a single ingot of n-type Ge ($\rho=6$ to 7 ohm-cm). The sample thicknesses have been 20 to 40 mils, and they have been oriented so that the magnetic field is perpendicular to the (100) plane. Sample diameters have been about 0.125 inch.



(a)

(b)

Fig. 11—Examples of poor (a) and excellent (b) germanium circulator samples, with bonded gold wires.

The first sample was made by thermo-compression bonding[3] a gold wire to each of the six faces of an approximately hexagonal slab. The spacing of the terminals was so irregular that the results were very poor, with no parallel network.

Subsequent circulators were made from samples cut with an ultrasonic cutter. These samples have six projections which lend themselves to being electroplated with antimony-doped gold without getting any plating on any of the rest of the surface. This plating was then alloyed into the pips and small balled gold wires were thermo-compression bonded to the centers of the plated surfaces. This technique produced quite symmetrical circulators.

The results of using these two different methods are shown in Fig. 11.

Most of the experimental work has been done in a four-inch Varian electromagnet, but the device will

[3] O. L. Anderson, H. Christensen, and P. Andreatch, "Technique for connecting electrical leads to semiconductors," *J. Appl. Phys.*, vol. 28, p. 923; August, 1957.

ultimately be used with a relatively small permanent magnet.

Practical Limitations

Assuming external stray reactances can be made negligibly small if proper care is taken in the design and use of the device, the speed of response will be limited only by the properties of the semiconductor material. That is, all the circulator admittances are pure conductances; consequently, the loads should also be pure conductances. This accounts for the fact that, theoretically, the circulator should function properly over the entire frequency range from dc up to the dielectric relaxation frequency of the material, which is of the order of a few kmc in the Ge used in these experiments. Unfortunately, the sample geometry requires that all leads be fairly close to one another, so radiation and pickup will undoubtedly place a lower practical limit on the usable frequency range. Skin effect will probably become important at high frequencies, also. These practical limitations have not yet been determined.

If the Hall effect circulator is used in an atmosphere whose temperature is not kept constant, both μ_n (the electron mobility) and B, (and thus θ_H), will depend somewhat on temperature. It is hoped that the following condition can be obtained:

$$\frac{\partial \mu_n}{\partial T} \approx - \frac{\partial B}{\partial T} \tag{23}$$

so that

$$\frac{\partial \theta}{\partial T} \approx \frac{\partial(\mu_n B)}{\partial T} \approx 0 \tag{24}$$

where T is temperature.

Impedance levels from 200 to 600 ohms have been attained on various samples, and it should be fairly simple to extend this range to 10 to 5000 ohms, still using only n-type Ge. With InAs and InSb, the low end of this range could be brought down to about 0.05 ohm. Of course, to be usable such a circulator would require extremely low resistance contacts. The use of high-mobility materials (*e.g.*, InSb, InAs, HgSe) would permit one to operate with smaller permanent magnets; however, circulators made of such materials would have extremely low impedance levels (perhaps on the order of the one-ohm) and would not be useful in most circuits.

One tremendous advantage this device has over most semiconductor devices is that it has no junctions. Because of this, its lifetime will probably depend only on the stability of the permanent magnet used. It requires no special atmosphere for good operation. If extremely thin samples were used, surface effects would play a more important role and special housings might be necessary.

The contacts, of course, must be ohmic. If not, their resistance will vary with current level, and so will the admittance of the device. As a consequence, the circulator would be balanced only for one particular signal level.

Concluding Remarks

It is possible to make three-port nonreciprocal Hall effect devices which will circulate dc and ac signals either in a clockwise or counterclockwise sense. With the aid of simple six-resistor networks (or even simpler three-resistor networks) the nine short-circuit admittance parameters of the circulator can be adjusted in a calculable manner. Incidentally, the corresponding parameters of any six-terminal circuit or black box could be adjusted in a similar manner without changing anything inside the three-port box. Modified forms of the parallel circuits might prove useful in conjunction with circuits with more or less than three ports. It should be remembered that the admittance equations would be equally valid for the case of complex admittances. In Hall effect circulators, an almost unlimited number of tricks must be possible. We have seen that such circulators can be made to operate at any desired impedance level, can be made to operate with any desired magnetic field, and can be made to introduce no loss—or even gain—if one uses negative resistances.

Circulators could be used in place of the hybrid coils in two-way, two-wire amplifiers. However, the 17 db forward loss will undoubtedly prevent them from being used in this way except, possibly, at frequencies above and below the usable frequency range of hybrid coils.

A circulator permits a transmitter and receiver to use a common antenna simultaneously if the circulator's losses are acceptable.

A circulator's sensitivity to variations in load impedance suggests its possible use as an impedance measuring device. In this case, it would measure the power reflected from a load, giving an indication of amount of impedance mismatch.

Probably the most important applications of circulators have yet to be devised. They have not made themselves evident previously because of the absence of circulators for "wire" frequencies. Whoever discovers such an application must not be discouraged by the circulator's high forward loss and its possible sensitivity to ambient temperature variations, and he must take advantage of the circulator's unique property: nonreciprocity.

Appendix

A Lower Bound on Forward Loss

At $B = B_0$, L_F is about 17.5 db. As the flux density is increased beyond B_0, L_F is decreased and asymptotically approaches a value corresponding to the maximum Hall angle which is theoretically possible.

To evaluate this minimum L_F, it will be advantageous to start with the open-circuit impedances (rather than the short-circuit admittances). When a voltage is applied to one of the three-terminal pairs with the other terminals open-circuited, the magnitudes of the two out-

ut voltages will approach (but not exceed) the input oltage as $B \rightarrow \infty$. Thus, let us assume

$$\lim_{B\rightarrow\infty} |z_R| = \lim_{B\rightarrow\infty} |z_F| = \lim_{B\rightarrow\infty} |z_S| = z_\infty \quad (25)$$

here z_R, z_F, and z_S are the three open-circuit imped-nces (defined in a manner analogous to that used for y_R, y_F, and y_S) of a symmetrical circulator. Measured alues of these impedances are plotted against B for an -type Ge circulator in Fig. 12. It can be seen from this igure that, in the limit of very large B, (25) leads to the esult that

$$z_R = -z_F = z_\infty. \quad (26)$$

Vith these impedance values, a simple calculation yields he following admittances:

$$y_S = y_F = \frac{1}{2z_\infty} \quad (27)$$

$$y_R = 0$$

which, when normalized, yield

$$U_S = \frac{y_S}{y_S} = U_F = 1$$

$$U_R = 0. \quad (28)$$

Substituting (27) and (28) in (17) gives for the condition for circulator action

$$k = 1 - U_L + \sqrt{(1 - U_L)^2 + 4}. \quad (29)$$

Requiring that the load admittance equal the input admittance gives

$$U_L = 1 + k + \frac{\left(1+\frac{k}{2}\right)^3 + \left(\frac{k}{2}\right)^3 - k\left(1+\frac{k}{2}\right)(1+k+U_L)}{(1+k+U_L)^2 - \frac{k}{2}\left(1+\frac{k}{2}\right)}. \quad (30)$$

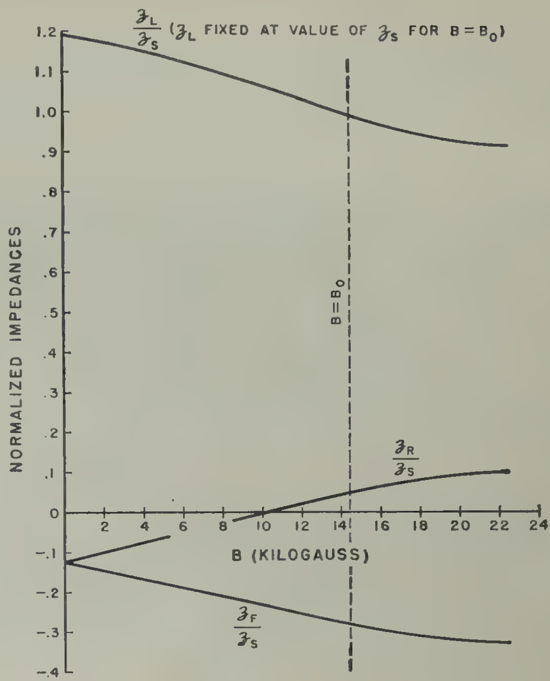


Fig. 12—Variation of open-circuit impedances with magnetic field.

Solving (29) and (30) simultaneously tells one that

$$U_L = 2.00$$

and

$$k = 1.24. \quad (31)$$

Using the values given in (28) and (31), one calculates the lowest possible forward loss for this type circulator to be 8.4 db.

ACKNOWLEDGMENT

The author wishes to express his appreciation for W. A. Taylor's assistance in preparing samples and making measurements.

Theory of the Crestatron: A Forward-Wave Amplifier*

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Summary—In the past a considerable amount of experimental evidence presented by various workers has indicated that gain apparently occurs in a traveling-wave type device even though the voltage may be so high that growing waves cease to exist. This means, in terms of Pierce's traveling-wave tube theory, that the growth constant of the growing wave x_1 is zero in this regime of operation. A theory explaining this phenomenon has been worked out and both small-signal and large-signal calculations have been carried out to investigate the characteristics of this type of operation. The gain occurs in this region due to a beating effect produced between the three small-signal forward waves described in traveling-wave tube theory as they travel along the RF structure. The maximum achievable gain is determined by the injection velocity and not by the length of the tube as in the normal case. Based on the above theory a device named the *Crestatron*, which utilizes this new mode of operation, has been built and tested to verify the theory, and it has been found that moderate gain (10–20 db) and high operating efficiency coupled with a very short length (4–6 wavelengths) characterize this mode of operation.

List of Symbols

$b = u_0 - v_0/Cv_0$ = injection velocity parameter
C = gain parameter
d = loss parameter
F = helix impedance reduction factor
f = frequency
I_0 = dc stream current
$\bar{\imath}$ = RF convection current in the beam
K_s = sheath-helix impedance
K_s' = normalized sheath-helix impedance
N = structure length in wavelengths
QC = space-charge parameter
u_0 = stream velocity
V = input RF voltage amplitude
V_{ci} = circuit component of wave amplitude
V_i = wave-voltage amplitude
V_0 = dc stream voltage
$\bar{v}$ = RF velocity in the beam
v_0 = circuit characteristic wave-phase velocity
z = distance measured from the input
$\beta = \omega/v$ = wave-phase constant
$\beta_e = \omega/u_0$ = stream-phase constant
Γ = wave-propagation constant in the presence of the beam
$\delta_i = x_i + jy_i$ = wave-incremental propagation constant
η = charge-to-mass ratio of the electron
η_s = saturation efficiency
$\theta = 2\pi CN$ = radian length of the tube
λ_g = guide wavelength
λ_s = stream wavelength
ω = angular frequency.

* Original manuscript received by the IRE, October 31, 1958. Revised manuscript received, December 22, 1958. The work was supported by the Rome Air Dev. Center, Griffiss Air Force Base, Rome, N. Y., under Contract No. AF 30(602)-1596 and AF 30(602)-1845.
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Introduction

It is well known that the operation of backward-wave devices, both O-type and M-type, depends on an interference phenomenon resulting from the beating between waves propagating along an RF structure. It can also be shown that in forward-wave devices, such as the traveling-wave amplifier, gain can occur due to the beating of waves traveling on the RF structure, providing the length is correct. The maximum achievable gain for any given set of circuit and operating parameters is determined by the relative injection velocity b rather than by the length of the tube as in the case of the conventional traveling-wave amplifier.

In order to clarify the principles and verify the theory proposed, a device, called the Crestatron, was designed and built to operate on the beating-wave principle. Measurements of gain on short-length tubes have indicated a substantial agreement with the theory. Theoretical calculations have also been correlated with measurements taken at other laboratories and very excellent agreement is obtained.

A voltage gain is produced in the Crestatron by a beating between the three forward waves propagating on the slow-wave structure. The device is operated with a beam velocity such that $b > b_{x_1=0}$ and, hence, there are no growing waves. The gain is achieved by adjusting the tube length to the proper value such that all the waves, one of which is out of phase with the others at the input, add together to give an RF voltage greater than that at the input. Hence there is not voltage gain in the sense of that produced by growing waves, but there is gain if the tube is considered to be a two-port network. The amount of gain is determined by the value of b and decreases as the beam velocity is raised above that at which growing waves cease to exist. The gain characteristics of the Crestatron can be calculated from the small-signal theory and these calculations also give information on the CN bandwidth ($\theta = 2\pi CN$) achievable in the device.

The large-signal theory of traveling-wave amplifiers[1] has been used to evaluate the nonlinear performance of the phenomenon as represented by the Crestatron in terms of the achievable gain and the expected operating efficiency.

Ample experimental evidence[2] has already been offered in this and other laboratories to verify the theory and indicate that high efficiency is obtained with short-

[1] J. E. Rowe, "A large-signal analysis of the traveling-wave amplifier: theory and general results," IRE Trans. on Electron Devices, vol. ED-3, pp. 39–56; January, 1956.
[2] J. J. Caldwell and O. L. Hoch, "Large-signal behavior of high power traveling-wave amplifiers," IRE Trans. on Electron Devices, vol. ED-3, pp. 6–18; January, 1956.

ength and moderate gains. During the course of this work it was learned that Mourier and Sugai had shown that a similar type of operation is possible with the forward-wave magnetron amplifier.[3] They treated the special case of small C, no space-charge and zero-circuit loss. These conditions in the magnetron amplifier give rise to two forward-traveling waves which under certain conditions can beat with one another to produce gain.

Derivation of the Gain Equation

The three forward waves propagating on a traveling-wave tube RF structure are known to vary in the following manner:[4]

$$e^{-\Gamma z} = e^{-j\beta z}\cdot e^{\beta C\delta z}, \tag{1}$$

where β is the wave phase constant and is equal to β_e at synchronism. The normalized voltage V_z/V at any point along the RF structure may be written in terms of the amplitudes of the three waves, neglecting space charge, as

$$\frac{V_z}{V} = e^{-j(\theta/C)}\left[\frac{V_1}{V}e^{\delta_1\theta} + \frac{V_2}{V}e^{\delta_2\theta} + \frac{V_3}{V}e^{\delta_3\theta}\right] \tag{2}$$

where

$\delta_i = x_i + jy_i$, $i = 1, 2, 3$,
V_i/V = normalized voltage amplitude of each wave,
$\theta \triangleq \beta C z = 2\pi CN$,
N = structure length in wavelengths, and
C = gain parameter.

From the small-signal theory of the traveling-wave tube, the following expressions for the RF convection current and velocity are obtained by retaining terms proportional to C:

$$\frac{(1+jC\delta_1)}{\delta_1}V_1 + \frac{(1+jC\delta_2)}{\delta_2}V_2 + \frac{(1+jC\delta_3)}{\delta_3}V_3 = \left(\frac{ju_0C}{\eta}\right)\tilde{v} \tag{3}$$

and

$$\frac{(1+jC\delta_1)}{\delta_1^2}V_1 + \frac{(1+jC\delta_2)}{\delta_2^2}V_2 + \frac{(1+jC\delta_3)}{\delta_3^2}V_3 = \left(\frac{-2V_0C^2}{I_0}\right)\tilde{\imath}, \tag{4}$$

where

$\eta = e/m$, charge-to-mass ratio of the electron,
$\tilde{v}$ = RF velocity in the stream,
$\tilde{\imath}$ = RF convection current in the stream,
V_0 = dc stream voltage, and
I_0 = dc stream current.

[3] G. Mourier and I. Sugai, private communication.
[4] J. R. Pierce, "Traveling-Wave Tubes," D. Van Nostrand Co., Inc., New York, N. Y.; 1950.

If an unmodulated stream is injected at $z=0$ and an RF signal level V is applied to the RF structure at that point, the boundary conditions require that the right-hand sides of (3) and (4) be zero and that

$$V = V_1 + V_2 + V_3. \tag{5}$$

Eqs. (3), (4) and (5) may be solved simultaneously to give the normalized amplitudes of the individual waves. The general result is

$$\frac{V_i}{V} = \left[1 + \frac{1+jC\delta_i}{1+jC\delta_{i+1}}\left(\frac{\delta_{i+1}}{\delta_i}\right)^2 \frac{\delta_{i+2}-\delta_i}{\delta_{i+1}-\delta_{i+2}} + \frac{1+jC\delta_i}{1+jC\delta_{i+2}}\left(\frac{\delta_{i+2}}{\delta_i}\right)^2 \frac{\delta_i-\delta_{i+1}}{\delta_{i+1}-\delta_{i+2}}\right]^{-1}, \tag{6}$$

where $\delta_i \equiv \delta_{i+3}$. Eq. (6) gives the total voltage associated with each wave and in the absence of space charge also gives the circuit voltage. The effect of passive modes or space charge is to reduce the circuit voltage from the value predicted by (6). The ratio of the circuit voltage to the total voltage is found from the ratio of the second term on the right-hand side of the following familiar quartic determinantal equation to the total right-hand side[5] of (6):

$$\delta^2 = \frac{(1+jC\delta)^2[1+C(b-jd)]}{\left[-b+jd+j\delta+C\left(jbd-\frac{b^2}{2}+\frac{d^2}{2}-\frac{\delta^2}{2}\right)\right]} - 4QC(1+jC\delta)^2. \tag{7}$$

The general result for the circuit component of voltage is

$$\frac{V_{ci}}{V_i} = 1 + 4QC\frac{(1+jC\delta_i)^2}{\delta_i^2}\cdot \tag{8}$$

Eqs. (6) and (8) have been obtained by Brewer and Birdsall.[6]

Thus the voltage along the RF structure, including the effects of finite C and space charge QC, generally may be written as

$$\frac{V_z}{V} = e^{-j(\theta/C)}\sum_{i=1}^{3}\left(\frac{V_i}{V}\right)\left(\frac{V_{ci}}{V_i}\right)e\delta_i{}^{\theta}, \tag{9}$$

where $\delta_i \equiv \delta_{i+3}$. When C is small and the effect of passive modes or space charge is negligible, (6) and (8) reduce to the following familiar form:

$$\frac{V_i}{V} = \frac{\delta_i^2}{(\delta_i-\delta_{i+1})(\delta_i-\delta_{i+2})} \tag{10}$$

and

$$\frac{V_{ci}}{V_i} = 1. \tag{11}$$

[5] *Ibid.*, p. 113, eq. (7.13) with corrections.
[6] G. R. Brewer and C. K. Birdsall, "Normalized Propagation Constants for a Traveling-Wave Tube for Finite Values of C," Tech. Memo. No. 331, Hughes Res. and Dev. Labs., Culver City, Calif.; October, 1953.

The voltage gain is written as

$$G_{\mathrm{db}} = 10 \log \left(\frac{V_z V_z^*}{VV^*}\right) = 10 \log \left|\left(\frac{V_z}{V}\right)^2\right|. \qquad (12)$$

It is important to note that, unlike conventional standing waves on a transmission line, each wave sees the RF structure characteristic impedance at all points along the line.

The gain that occurs when the velocity parameter b is greater than that for which the growth constant of the growing wave is zero is, as mentioned before, due to a beating effect between the three small-signal waves which are set up at the input and propagate along the RF structure. The energy taken from the stream in this type of operation goes into setting up the three forward-circuit waves at the input, and then these waves propagate with different velocities and constant amplitude along the RF structure. The gain occurs due to the eventual adding in phase of the two larger waves. The interaction is primarily between the circuit wave and the slow space-charge wave. It will be seen later that the fast space-charge wave is excited to a negligible extent. This is the same basic mode of operation as in the backward-wave device of both the traveling-wave tube and crossed-field types. As the three waves travel along the structure, the phase relationship between the RF current in the beam and the RF field on the circuit changes, and at certain points along the circuit the phase is such that energy is transferred to the circuit. At the same time there are certain regions along the tube where the phase relationship between the beam current and the circuit field is such that energy is fed from the circuit back to the beam. The realizable gain in this mode of operation depends upon the relative injection velocity b for any given set of tube parameters rather than on the length as in the case of the conventional traveling-wave tube.

A forward-wave embodiment of the beating phenomenon, such as the Crestatron, is inherently more efficient than the backward-wave devices which operate on the same beating principle because the modulation in the stream and the field on the circuit producing the modulation travel in the same direction, whereas in the backward-wave device the modulation in the beam and the circuit field travel in opposite directions. In the backward-wave device the circuit field is strongest where the modulation is weakest and vice versa.

Mathematically speaking, all the energy is abstracted from the beam at the input since in satisfying the boundary conditions energy is put into setting up the three circuit waves. Then the circuit length is simply adjusted so that at the end of the tube the wave energies all add in phase. It should be recalled that in this region of operation the propagation constants are purely imaginary, giving rise to real voltage amplitudes, and at the input one voltage component is 180° out of phase with the other two.

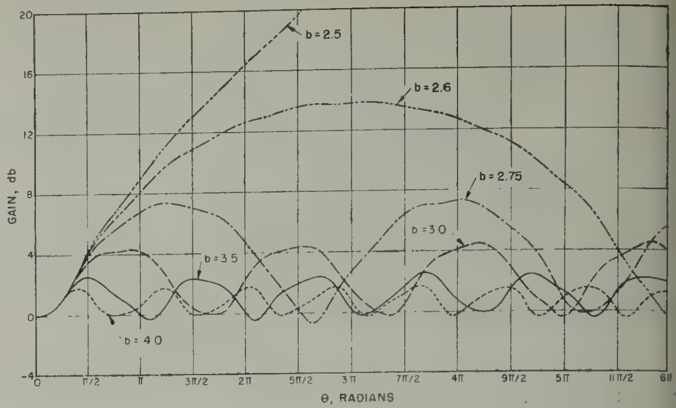

Fig. 1—Gain vs length ($C=0.1$, $QC=0.25$, $d=0$, $b_{x_1=0}=2.57$).

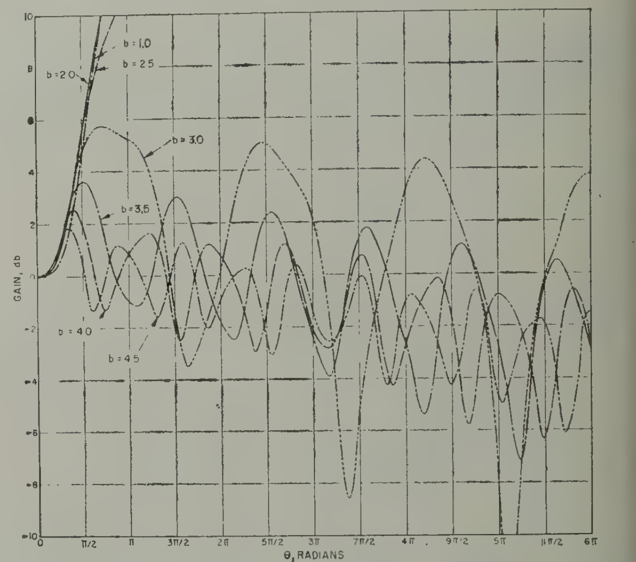

Fig. 2—Gain vs length ($C=0.2$, $QC=0.125$, $d=0.025$, $b_{x_1=0}>3.0$).

Small-Signal Gain

It was pointed out above that the normalized voltage amplitudes are purely real since the propagation constants are purely imaginary when $b>b_{x_1=0}$. For fixed values of C, QC, and d the gain may be calculated from (9) and (12) as a function of θ for particular values of b. Typical gain curves are shown in Figs. 1 and 2. The above gain equations are valid for all values of b, and it is seen from the figures that the normal exponential gain is obtained when $x_1 \neq 0$. The gain curves were plotted over a range of 6π radians to indicate their repetitive nature. In Fig. 2 it is seen that the effect of circuit loss is to reduce the gain, particularly for large θ.

The results plotted in Fig. 2 do not show the large negative dip in the gain curve for $b=0$ that Brewer and Birdsall[7] found. The fact that they assumed a $QC=0.25$ whereas the above data are for $QC=0.125$ is of little consequence to this discrepancy. It is believed that their results are incorrect, since to compute gain they used the

[7] G. R. Brewer and C. K. Birdsall, "Traveling-wave tube propagation constants," IRE Trans. on Electron Devices, vol. ED-4, pp. 140–144; April, 1957 (Fig. 2).

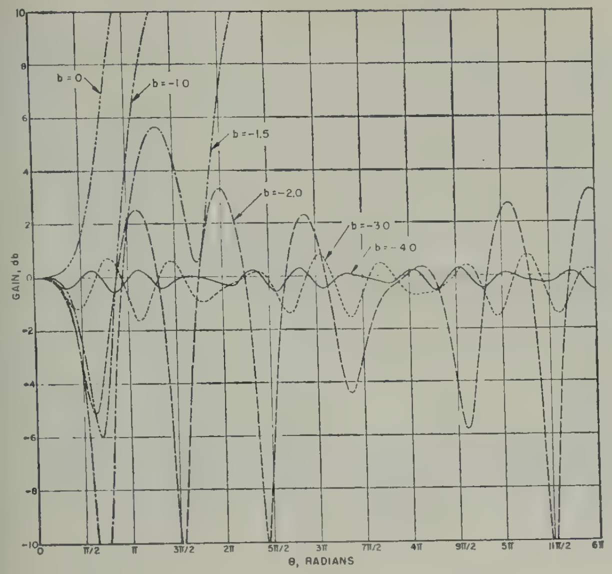

Fig. 3—Gain vs length ($C=0.1$, $QC=0.125$, $d=0$, $b_{x_1=0}=2.34$).

small-C equations, which are not valid when $C=0.2$ and $QC=0.25$. It should be pointed out, as is predicted here, that only very small dips in the gain curves were noted in the large-signal calculations. Large negative dips in the gain seem to occur at $b=-1$ for most values of C and QC and for positive values of b when d is large.

The gain curves of Fig. 3 show that gain is also achieved for negative values of b such that $x_1=0$. Negative values of b correspond to operating voltages less than the synchronous voltage, and the resulting gains are less than those obtained with large values of b.

When the loss parameter is zero all maxima of the gain curve are approximately equal (the slight variations and lack of periodicity will be explained later), but when $d\neq 0$ the first maximum will be highest and subsequent peaks will generally be lower. This merely emphasizes the fact that when circuit loss is significant the tube length should be chosen so as to operate on the first maximum of the gain curve. Because of the relatively low gain of the Crestatron a large attenuator like those used in ordinary traveling-wave amplifiers is not necessary.

It is seen from the gain curves that for fixed b and variable θ the curves are nonperiodic within the interval 6π and exhibit periodic undulations in amplitude. The explanation of this phenomenon for voltage amplitude vs distance is contained in (9). The lack of periodicity with 2π and the undulating peak amplitudes are a result of the product of exp $(-j\theta/C)$ and exp $(\delta_i\theta)$, where each represents a vector rotating about the origin as a function of θ. The rate of rotation of the first vector is related to $1/C$, which is typically between 5 and 20, and the rate of rotation of the second is determined by δ_i, which varies between 1 and 3. The first, then, perturbs the second as a modulation, and the fact that δ_i is nonintegral in general means that it is not periodic with 2π. It may be that θ must travel through $2m\pi$ radians with m very large before a periodicity is apparent, if ever.

The above phenomenon indicates that there is something more than a simple beating effect occurring between the waves. In fact there is a continual slipping of one wave with respect to the others along the RF structure. This process accounts for the fact that there is or may be a net interchange of energy between the beam and the circuit in a device of infinite length. It is interesting to examine the condition necessary for the gain curve to be periodic with period $2m\pi$ and also the condition required for the peak amplitudes to be constant. In fact these two conditions would result in a gain curve that is exactly reproduced every 2π radians.

For V_z/V in (9) when b is held fixed, the phase condition is determined by the product

$$e^{-j(\theta/C)}\cdot e^{\delta_i\theta} = e^{j\theta(y_i-1/C)}, \tag{13}$$

since $x_1=0$. Thus in order for the phase to be periodic with $2m\pi$,

$$n\left(y_i-\frac{1}{C}\right) = 2m\pi. \tag{14}$$

Solving for y_i gives

$$y_i = \frac{2m\pi}{n}+\frac{1}{C}, \tag{15}$$

where n and m are independent integers. Thus for phase periodicity the value of y_i must be equal to a constant plus some integral or subintegral multiple of 2π. When $C\to 0$, $y_i=1/C$ and all the waves travel at approximately the same rate, they maintain the same relative phase positions with respect to one another. This results in a constant-amplitude vector rotating about the origin with a period of 2π radians. Thus the gain curve will be periodic in amplitude also.

Under very restricted conditions a simplified expression for the gain of the tube may be obtained from (9). When $C\to 0$, $QC=0$, $d=0$, and b is sufficiently large that the propagation constants are purely imaginary, it can be determined that the δ_i's are given approximately by

$$\delta_1 \approx -j/b^{1/2},$$

$$\delta_2 \approx -jb,$$

and

$$\delta_3 \approx j/b^{1/2}. \tag{16}$$

Substitution of (16) into (9) yields for the gain, after some simplification,

$$\left|\frac{V_z}{V}\right|^2 = \left(\frac{1}{1-b^3}\right)^2\Bigg[1+b^6+(b^3-1)\sin^2\frac{\theta}{b^{1/2}} - 2b^3\left\{\cos\theta b\cos\frac{\theta}{b^{1/2}}+b^{3/2}\sin\theta b\sin\frac{\theta}{b^{1/2}}\right\}\Bigg]. \tag{17}$$

Since b is usually greater than 2.5, (17) can be simplified further to

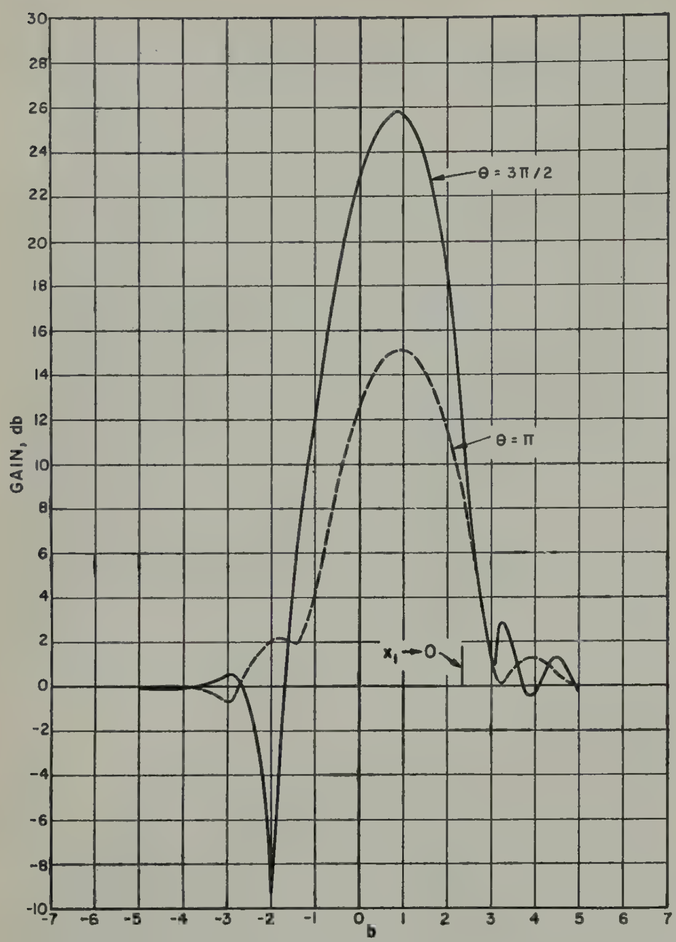

Fig. 4—Gain vs injection velocity for fixed tube length ($C=0.1$, $QC=0.125$, $d=0$).

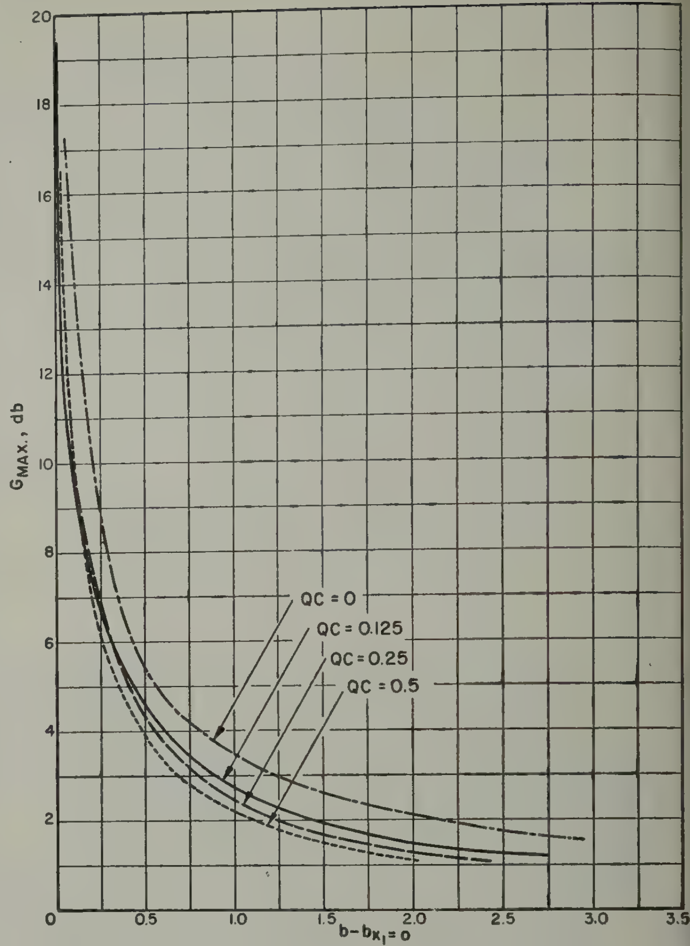

Fig. 5—Maximum gain vs $b-b_{x_1=0}$ with space charge as the parameter ($C=0.1$, $d=0$).

$$\left|\frac{V_z}{V}\right|^2 = 1 + \frac{1}{b^3}\left[\sin^2\frac{\theta}{b^{1/2}} - 2\cos\theta b\cos\frac{\theta}{b^{1/2}} + b^{3/2}\sin\theta b\sin\frac{\theta}{b^{1/2}}\right]. \quad (18)$$

A useful expression for predicting the value of CN at the first maximum of the gain curve may be obtained by differentiating (17) with respect to θ and setting the result equal to zero. In this way

$$\sin\frac{\theta}{b^{1/2}}\left[\cos\frac{\theta}{b^{1/2}} - b^3\cos\theta b\right] = 0. \quad (19)$$

Thus maxima and minima in the gain occur for

$$\sin\frac{\theta}{b^{1/2}} = 0$$

or

$$CN = \frac{nb^{1/2}}{2} \qquad n = 0, 1, 2, \cdots. \quad (20)$$

The other condition is

$$\cos\theta b = \frac{\cos\frac{\theta}{b^{1/2}}}{b^3} \approx 0 \text{ for large } b$$

or

$$CN = \frac{2n+1}{4b}, \qquad n = 0, 1, 2, \cdots. \quad (21)$$

Eq. (21) predicts quite accurately the distance to the first maximum in the gain curve when $n=1$. Thus

$$CN \text{ to first maximum} = \frac{0.75}{b}. \quad (22)$$

Eqs. (20) and (21) coupled with information on the second derivative of the gain curve may be used to determine subsequent maxima in the gain vs θ curve.

In addition to gain vs length curves for fixed b, a set of gain curves may be obtained for fixed tube length with variable voltage or b. Typical curves of this type are shown in Fig. 4. The main hump in the gain curve vs b is of course due to exponential gain, and the other lower peaks correspond to gain through beating waves. When $C\to 0$, $QC=0$ and $d=0$ the gain vs b curve will

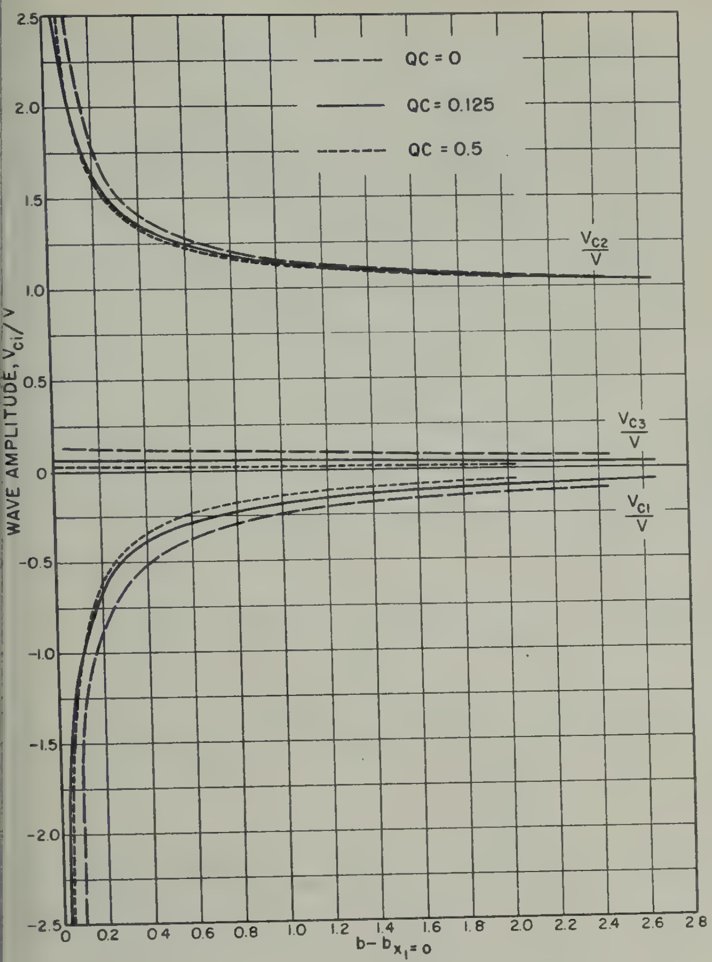

Fig. 6—Normalized amplitudes of the wave voltages vs injection velocity ($C=0.1$, $d=0$).

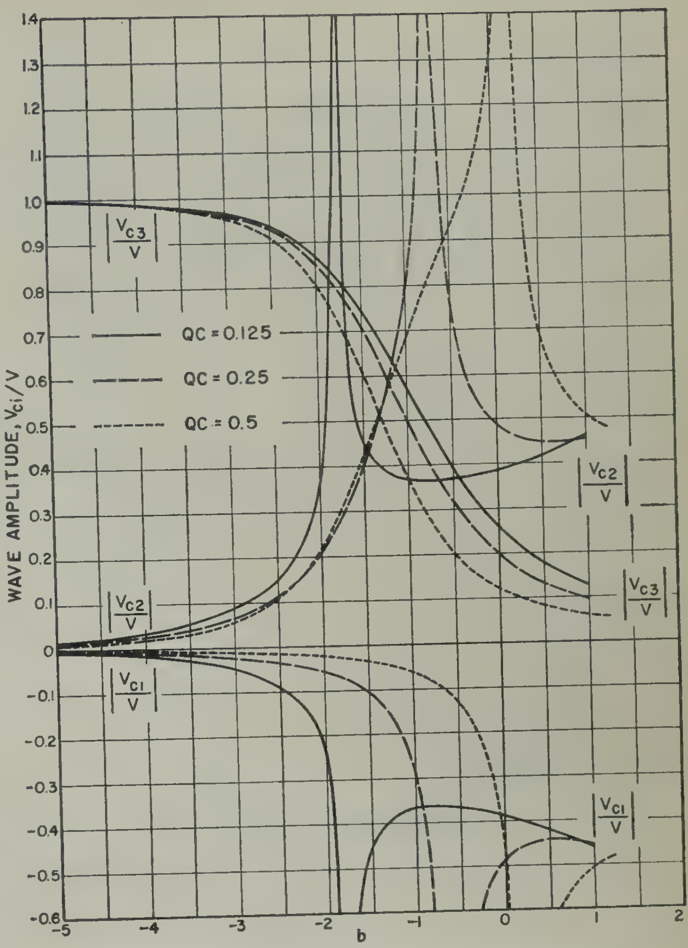

Fig. 7—Normalized amplitudes of the wave voltages vs injection velocity ($C=0.1$, $d=0$).

be symmetrical with respect to b since maximum exponential gain occurs at synchronism.

Bandwidth

The bandwidth of a device operating on the beating-wave principle may be determined from the gain curves shown in Figs. 1 and 2. The CN bandwidth ($\theta=2\pi CN$) can be as large as ± 50 per cent between 3-db points on the gain curve. If the impedance of the structure remained constant over this range, then the device would have a frequency bandwidth of 3:1. In general, impedance variations will limit this figure to some lower value. Experimental Crestatrons have been built with nearly an octave bandwidth at 3 kmc.

Maximum Gain vs C and QC

It was pointed out earlier that the voltages are purely real for $b>b_{x_1=0}$, and at the input one voltage is 180° out of phase with the other two. Thus the maximum gain will occur when the three waves all add in phase. The value of b for which $x_1=0$ can be computed from the quartic determinantal equation. When $d\neq 0$, x_1 is not exactly zero anywhere but does drop to less than one per cent of its maximum value for large b.

Maximum-gain curves for various values of QC when $C=0.1$ are shown in Fig. 5. The gain decreases with increasing voltage and for very high voltage approaches zero asymptotically. As would be expected the gain increases with C and decreases as QC is increased. The effect of loss on the circuit also reduces the gain. The greatest gain occurs for a b which is equal to that for which growing waves cease to exist. There is a smooth transition from the region of exponentially growing waves to the region in which the gain is a result of the beating of the three waves.

The gain for large positive values of b is a result of the combining of V_{c1}/V and V_{c2}/V, since V_{c3}/V is usually negligible. The magnitudes of these normalized voltages for a typical range of parameters are shown in Fig. 6. For large values of b, it is seen from Fig. 6 that V_{c1}/V approaches zero and V_{c3}/V is still small, so that almost all the energy is in the second wave (circuit wave) V_{c2}/V. On the other hand, for negative values of b (voltage below synchronism), the third wave (fast space-charge wave) V_{c3}/V predominates, as shown in Fig. 7.

A typical plot of the distance to the first maximum of the gain curve is shown in Fig. 8 for several values of space charge. The theoretical relation (22) is also shown.

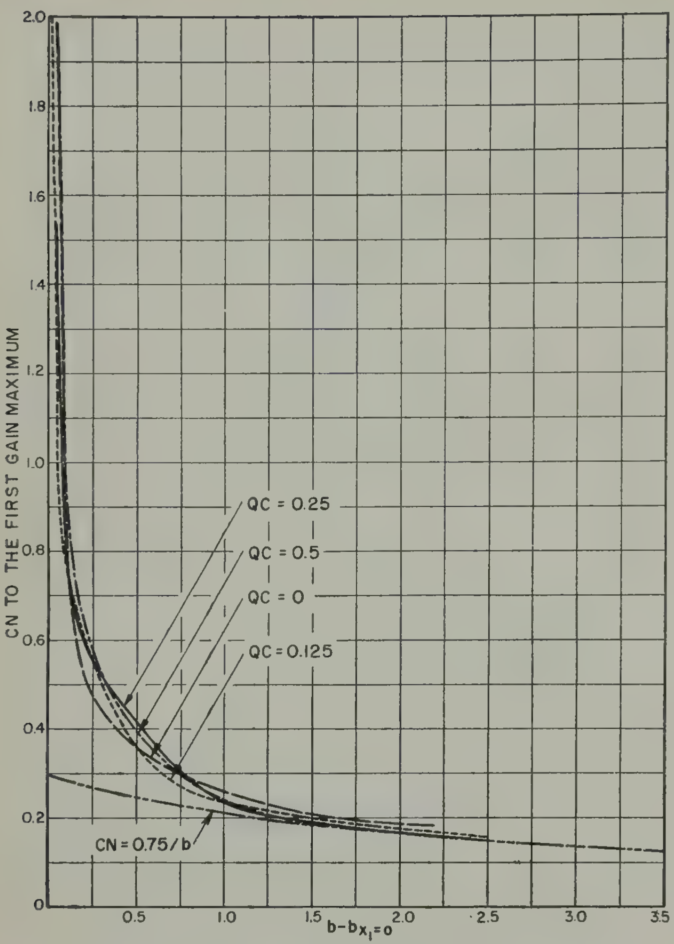

Fig. 8—CN to the first gain maximum vs injection velocity ($C=0.1$, $d=0$).

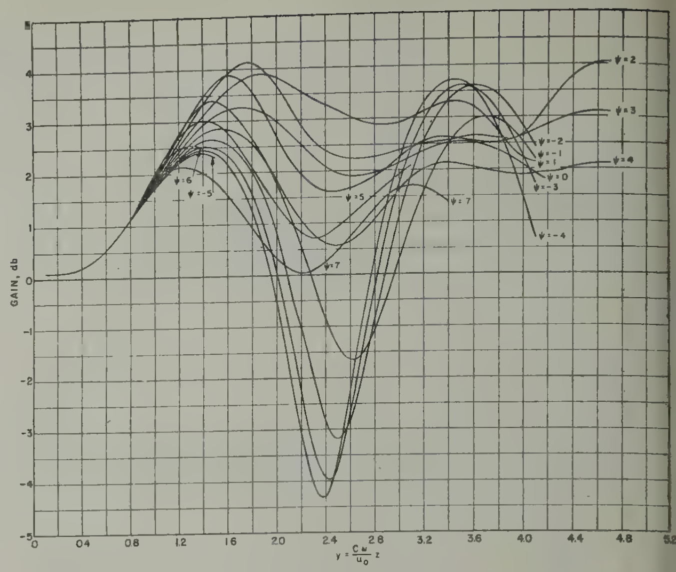

Fig. 9—Theoretical large-signal gain vs length ($C=0.1$, $QC=0.25$, $B=1.0$, $d=0$, $b=3.5$).

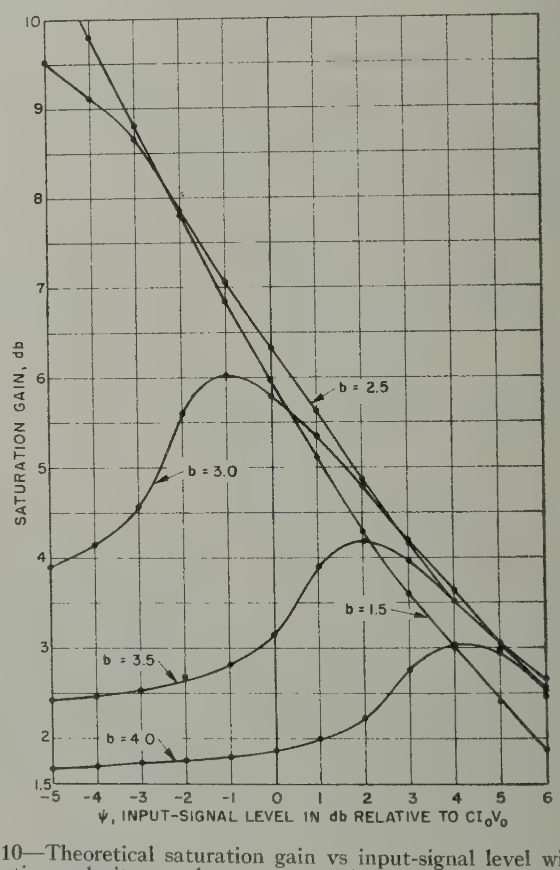

Fig. 10—Theoretical saturation gain vs input-signal level with injection velocity as the parameter ($C=0.1$, $QC=0.125$, $B=1$, $d=0$, $b_{x_1=0}=2.33$).

Large-Signal Performance

Saturation Gain

The large-signal performance of a beating-wave device has been evaluated using the large-signal theory of the traveling-wave amplifier.[1] Significant large-signal gain is found for large values of $b > b_{x_1=0}$ when the input-signal level, ψ, is appreciable compared to CI_0V_0. A typical gain curve is shown in Fig. 9 with the input-signal level to the RF structure as the parameter. A periodicity in the large-signal gain vs distance curves is found similar to that previously shown with the linear theory. The gain is seen to decrease as the drive level is increased, but the power output and efficiency increase as the drive level is increased.

The composite results of the large-signal calculations may be plotted in summary form to show several interesting facets of the nonlinear performance of the Crestatron. As seen in Fig. 10 the "saturation" gain as a function of the drive level to the RF structure goes through a maximum at a value of ψ dependent upon the value of b. It should be noted that the optimum length, *i.e.*, for maximum output, changes as the drive level changes. When the tube is operated at a b value which gives maximum small-signal gain or maximum saturation efficiency, an increase in the drive level results in a smooth transition to the beating-wave type of operation and the gain decreases smoothly. For very high drive levels the gain is low and relatively independent of the injection velocity. This relationship is also in line with that predicted from the linear theory. The value of CN

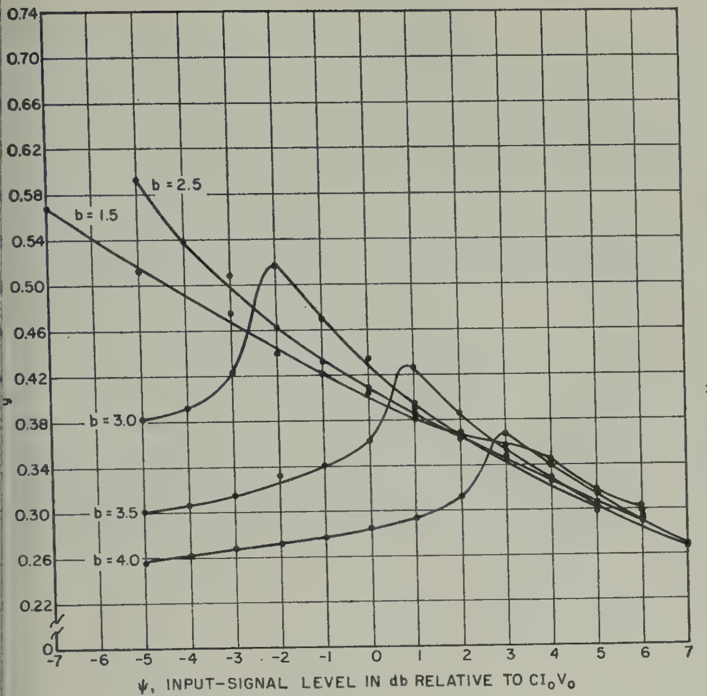

Fig. 11—Theoretical saturation length vs input-signal level with injection velocity as the parameter ($C=0.1$, $QC=0.125$, $B=1.0$, $d=0$, $b_{x_1=0}=2.33$).

to the first maximum of the gain curve can also be plotted as a function of both the drive level and the value of b as shown in Fig. 11.

A considerable amount of previously unexplained experimental information has been obtained by Caldwell and Hoch[2] and other workers which verifies the existence of gain in a forward-wave amplifier due to the beating between the waves propagating on the RF structure. A comparison of the calculated gains using the linear and nonlinear theories along with some data extracted from Caldwell and Hoch's paper is presented in Fig. 12. It is seen that the agreement is very good between the theoretical results and the experimental data. Similar experimental data have been obtained in this laboratory which substantiate the theory even more clearly. (It is planned to publish complete experimental data on a series of Crestatrons in a later paper.) Since the value of d was approximately 0.025 for the experimental data, it is difficult to choose a reference b where x_1 is nearly zero. A value of $b=2.9$ was used, but had a larger value been used the agreement would have been even better.

Caldwell and Hoch found in addition that the saturation power output increased as the drive level was increased, keeping the voltage constant. This phenomenon is also predictable from the nonlinear theory; a comparison is shown in Fig. 13. The amount of increase in the saturation power level for input-signal levels comparable to or greater than CI_0V_0 is some 4 db over that for very small input-signal levels. This increase in power output is probably due to the fact that this type of operation extracts the energy associated with all three circuit waves rather than only that of the growing wave as in the case of the conventional traveling-wave amplifier.

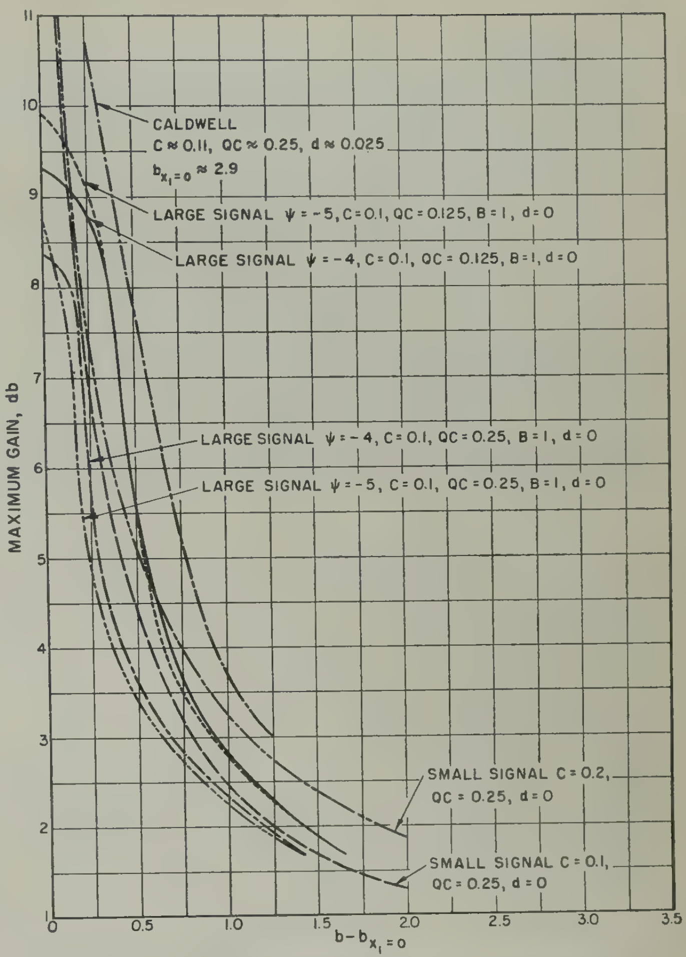

Fig. 12—Comparison of theoretical and experimental gain.

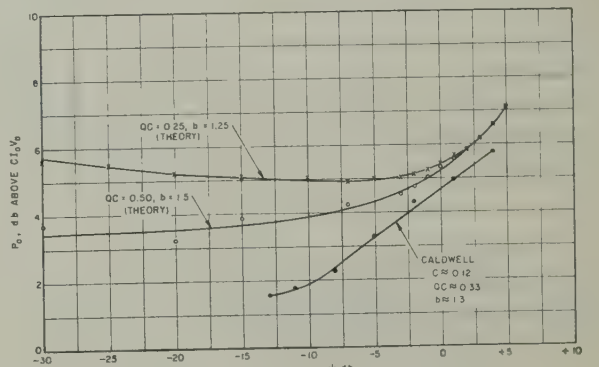

Fig. 13—Variation of saturation level with input-signal level ($C=0.1$, $B=1.0$, $d=0$).

Efficiency

The saturation efficiency also increases as the drive level is increased, resulting in a relatively high efficiency for the Crestatron. Efficiency curves as a function of the drive level and the injection velocity are shown in Fig. 14. It is seen that the saturation efficiency for high drive approaches a value relatively independ-

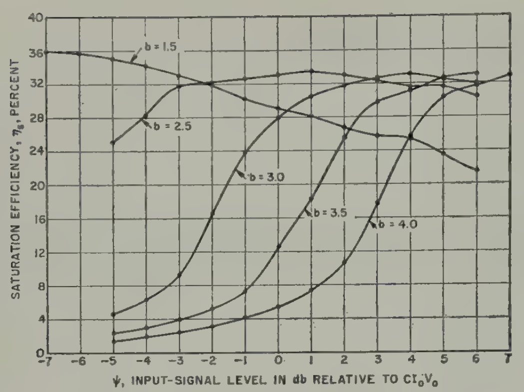

Fig. 14—Theoretical saturation efficiency vs input-signal level with injection velocity as the parameter ($C=0.1$, $QC=0.125$, $B=1.0$, $d=0$, $b_{x_1=0}=2.33$).

ent of the value of b in this type of device. In calculating the saturation efficiency of the Crestatron the input power is significant and, hence, one must consider only the energy conversion efficiency and subtract out the input-power level from the calculations. For values of b just slightly greater than that for which the wave growth constants become zero the efficiency is fairly constant as the drive level is varied. Continual increase in the drive power when b is adjusted for maximum saturation efficiency with small input-signal level results in a gradual decrease in the saturation efficiency.

The saturation efficiency is apparently higher at high values of drive when the voltage is adjusted for beating-wave operation than when the voltage is adjusted for maximum saturation output for a small input-signal level. This increase in efficiency is probably due in part to the fact that $1+Cb$ is approximately 1.3 or 1.4 in the Crestatron, as opposed to around 1.2 in the growing wave mode of operation and hence has considerably more energy to give up to the wave while slowing down.

Magnetic Field Requirement

The limiting stream perveance which can be transmitted through a cylindrical drift tube or helix due to space-charge spreading is given by[8]

$$\text{Perveance} = 38.6 \times 10^{-6}\left(\frac{d}{l}\right)^2, \tag{23}$$

where

d = helix diameter, and
l = helix length.

The following familiar relations for traveling-wave tubes are recalled:

$$CN_s = \left(\frac{FK_sI_0}{4V_0}\right)^{1/3}\frac{lf}{u_0}, \tag{24a}$$

$$u_0 = (2\eta V_0)^{1/2} = \lambda_s f, \tag{24b}$$

$$\frac{u_0}{v_0} = 1 + Cb, \tag{24c}$$

and

$$\gamma a = 2\pi a/\lambda_g. \tag{24d}$$

The helix impedance FK_s used in (24a) may be written as

$$FK_s = \frac{506FK_s'(1+Cb)}{V_0^{1/2}}, \tag{25}$$

where

F = the impedance reduction factor due to dielectric loading and space harmonics, and
K_s' = sheath-helix impedance as given by Pierce[9] in Fig. A6.5.

Combining (23)–(25) yields

$$CN_s = 0.054\frac{(FK_s')^{1/3}}{(1+Cb)^{2/3}}(\gamma a)\left(\frac{l}{d}\right)^{1/3}. \tag{26}$$

From (26) can be calculated the maximum length of structure which can be used when no magnetic focusing field is present and the stream is allowed to spread under the influence of space-charge forces. Of course, the stream must be injected in a convergent manner, aimed toward the center of the structure, and then allowed to expand, resulting in a divergent stream at the output end. Allowance will also have to be made for stream spreading due to the presence of RF fields on the structure. The following typical values of the parameters in (26) are used to evaluate CN_s:

$$\gamma a = 1.5$$
$$(FK_s')^{1/3} = 2.2$$
$$(1+Cb)^{2/3} = 1.3$$
$$\left(\frac{l}{d}\right)^{1/3} = 2.5.$$

Using the above values gives

$$CN_s = 0.34.$$

This value of CN_s is compatible with the lengths required for Crestatron operation. Hence under some conditions it may be possible to operate the Crestatron with little or no magnetic focusing field. Electrostatic focusing systems are sometimes possible depending upon the structure type.

Conclusion

A theoretical investigation of traveling-wave tube operation when $b > b_{x_1=0}$ has revealed that gain can be

[8] J. R. Pierce, "Theory and Design of Electron Beams," D. Van Nostrand Co., Inc., New York, N. Y., p. 151; 1954.

[9] J. R. Pierce, "Traveling-Wave Tubes," D. Van Nostrand Co., Inc., New York, N. Y.; 1950.

hieved due to a beating between the three small-gnal waves which are set up at the input of the ampli-r. These waves propagate along the RF structure with nstant amplitudes and different velocities and hence eat with one another. Gain is achieved by terminating e circuit at a point where these waves add in phase to ve an increased output. The gains achievable are mod-ate and increase with increasing C but decrease with creasing QC. The gain also decreases as the injection elocity as measured by b is increased.

Nonlinear calculations indicate the same type of be-avior as both the voltage and drive level are increased; nd it has been shown that the saturation power output ncreases with the drive level. The saturation efficiency s high for this type of operation and the CN bandwidth an be as large as ± 50 per cent.

In order to check the theory experimentally a beating-wave device called the Crestatron has been designed and tested. Measurements of gain indicate excellent agreement with the theory as do comparisons with data taken by other workers. Since the RF structure length of the Crestatron is extremely short it has been shown that it is theoretically possible to operate the device with little or no magnetic focusing field.

Acknowledgment

The author acknowledges the benefit of discussions during the course of this work with his associates in the Electron Physics Laboratory, University of Michigan, Ann Arbor, and in particular the assistance of Y. C. Lim and L. E. Stafford in solving the equations on a digital computer.

Correction

The IRE Standards Committee has requested that the list of Committee Personnel which appeared with the "IRE Standards on Audio Techniques: Definitions of Terms, 1958," on page 1928 of the December, 1958 issue of Proceedings, be corrected as follows.

Subcommittee on Audio Definitions

D. S. Dewire, *Chairman* 1956–1958
L. D. Runkle, *Chairman*, 1955–1956

H. W. Bibber 1957–1958
D. S. Dewire 1955–1956
W. E. Darnell 1955–1958
W. F. Dunklee 1955–1958
C. W. Frank 1955–1958
A. A. McGee 1956–1958
G. H. Grenier 1955–1956
L. D. Runkle 1956–1958
R. E. Yaeger 1955–1958

Committee on Audio Techniques

D. S. Dewire, *Chairman* 1958
Iden Kerney, *Chairman* 1956–1958
D. E. Maxwell, *Chairman* 1954–1956
A. P. Evans, *Vice-Chairman* 1958
D. S. Dewire, *Vice-Chairman* 1956–1958
Iden Kerney, *Vice-Chairman* 1954–1956

Theory and Experiments on Shot Noise in Silicon *P-N* Junction Diodes and Transistors*

B. SCHNEIDER† AND M. J. O. STRUTT†, FELLOW, IRE

Summary—**Experiments with silicon junction diodes and transistors have shown that previous theoretical expressions for the noise of such elements do not hold for silicon. New theoretical expressions are derived on the basis of recombination-generation in the depletion layer. These new expressions are satisfactorily checked by experiments in the case of low-level current injection. At high-level injection, however, deviations occur, for which no exact theory is known.**

Introduction

EXPERIMENTS have shown that the theoretical expressions of noise in *p-n* junction diodes and transistors coincide reasonably well with experimental values in the case of germanium at low current densities.[1,2] With silicon, however, significant deviations occur.[3] It is well known that the characteristic curves of silicon *p-n* diodes and transistors show deviations from those of the corresponding germanium types.[4,5] In the low current density region, theoretical expressions for these characteristic curves of silicon devices have been put forward,[4,6,7] which show satisfactory coincidence with experimental values. In this paper, new expressions for the shot noise of silicon-junction diodes and transistors are derived on the basis of the theory of Sah, Noyce, and Shockley.[4] It is then shown that these new expressions are in satisfactory agreement with experimental values.

Differential Admittance of *P-N* Junction Diodes

In germanium *p-n* junction diodes current flow is mainly caused by diffusion and is, to a good approximation, given by Shockley's equation

$$I_D = I_o\left[\exp\left(\frac{qV}{kT}\right) - 1\right]. \tag{1}$$

Here, I_o is the saturation current, q the magnitude of electron charge, V the applied voltage, k Boltzmann's constant, and T the absolute temperature (degrees Kelvin). The positive direction of currents and voltage is counted from the *p*-side to the *n*-side of the junction.

From (1) the real part of the differential admittance Y_D is found to be

$$\text{Re}\,(Y_D) = \frac{dI_D}{dV} = I_o\frac{q}{kT}\exp\left(\frac{qV}{kT}\right) = \frac{q}{kT}(I_D + I_o). \tag{2}$$

If the diode is biased in the forward direction and $qV > 3kT$, say, we obtain approximately

$$\text{Re}\,(Y_D) \approx \frac{q}{kT}I_D. \tag{3}$$

These equations are approximately valid in the case of low diode current densities. They break down, also, for germanium *p-n* junction diodes, if high-level current injection occurs, *i.e.*, for high current densities.[8]

In the case of silicon *p-n* junction diodes, considerable recombination and generation of carriers occurs in the depletion layer. As a result of this, a recombination-generation current flows through the junction diode, called I_R in this paper. According to Sah, Noyce and Shockley,[4] this current is given by the expression

$$I_R = A\frac{\pi}{2}\frac{n_i kTw}{\tau_o(V_D - V)}\exp\left(\frac{qV}{2kT}\right). \tag{4}$$

Here, A is the cross-section area of the junction, n_i is the density of holes or electrons in an intrinsic specimen of the material under consideration, w is the width of the depletion layer, V_D is the diffusion voltage (or contact voltage) of the *p-n* junction, and τ_o is the mean lifetime of electrons and holes (if the lifetimes of either are equal). The current I_R, which is for short called "recombination current," gives rise to an admittance Y_R, the real part of which is, by (4),

$$\text{Re}\,(Y_R) = \frac{qI_R}{2kT}\left[1 + \frac{kT}{q(V_D - V)}\right]. \tag{5}$$

If we apply a forward bias voltage V, the second term within square brackets is mostly small with respect to unity, and hence, by (5),

* Original manuscript received by the IRE, September 15, 1958; revised manuscript received, December 8. 1958. This work was made possible by grants from the Swiss Federal Fund for Scientific Research, and from the Swiss Federal Institute of Technology.

† Swiss Federal Institute of Technology, Zürich, Switzerland.

1 W. Guggenbuehl and M. J. O. Strutt, "Theory and experiments on shot noise in semiconductor junction diodes and transistors," Proc. IRE, vol. 45, pp. 839–854; May, 1957.

2 A. van der Ziel, "Noise in junction transistors," Proc. IRE, vol. 46, pp. 1019–1038; June, 1958.

3 B. Schneider, "Untersuchungen des Hochfrequenzrauschens von neueren Transistoren," (High frequency noise of modern transistors"), Communication Nachrichtentechnische Gesellschaft Session, Karlsruhe, Ger., September 24, 1957.

4 C. T. Sah, R. N. Noyce, and W. Shockley, "Carrier generation and recombination in *p-n* junctions and *p-n* junction characteristics," Proc. IRE, vol. 45, pp. 1228–1243; September, 1957.

5 J. L. Moll, "The evolution of the theory for the voltage-current characteristic of *p-n* junctions," Proc. IRE, vol. 46, pp. 1076–1082; June, 1958.

6 W. Shockley and W. T. Read, Jr., "Statistics of recombination of holes and electrons," *Phys. Rev.* vol. 87, pp. 835–842; September, 1952.

7 R. N. Hall, "Electron-hole recombination in germanium," *Phys. Rev.*, vol. 87, p. 387; July, 1952.

8 W. Guggenbuehl, "Theoretische Ueberlegungen zur physikalischen Begründung des Ersatzschaltbildes von Halbleiterdioden bei hohen Stromdichten," *Arch. elekt. Übertragung*, vol. 10, pp. 483–485; September, 1956.

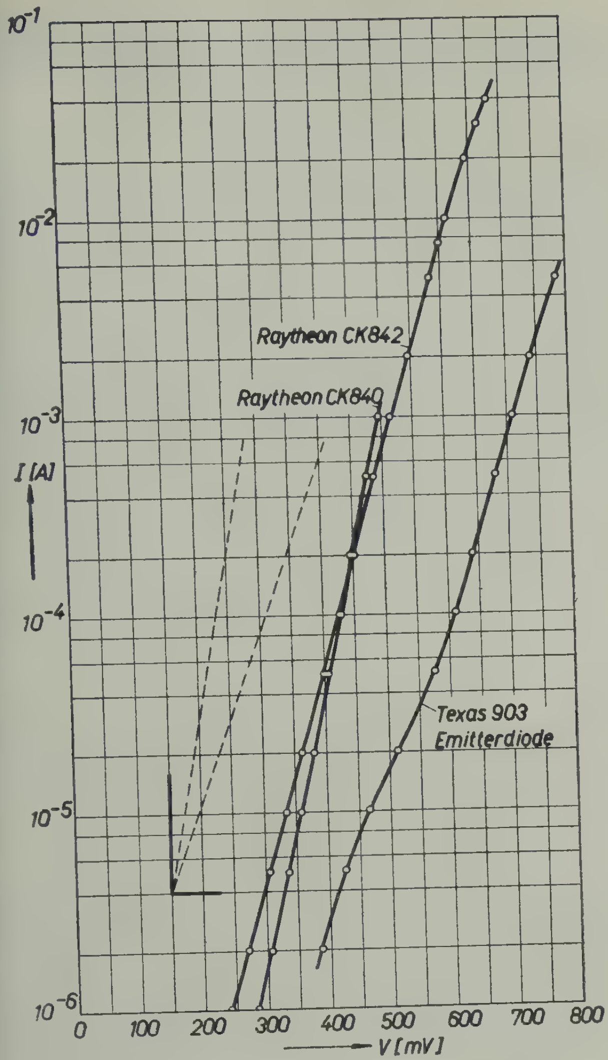

Fig. 1—Characteristic curves of diode current (vertical) vs diode voltage (horizontal) for three different silicon diodes (full curves), one of these being the emitter diode of a silicon transistor (Texas 903). The two broken curves show the slopes pertaining to pure diffusion (on the left) and to pure recombination (on the right).

$$\text{Re}\,(Y_R) \approx \frac{qI_R}{2kT}. \tag{6}$$

From (4) we cannot directly see why I_R is much more relevant in silicon p-n junction diodes than in germanium diodes. In order to see this, the recombination current I_R must be compared with the saturation currents I_o (which is the reverse current of a far negatively biased diode) in both cases. A simple argument[5] then shows that I_R is much more important, relatively to I_o with silicon than with germanium p-n junction diodes.

In general, a diffusion current, as well as a recombination current, will flow in a p-n junction diode, thus yielding a total current I,

$$I = I_D + I_R \tag{7}$$

The total differential admittance $Y = Y_D + Y_R$ has a real part, given by (3) and (6) for a forward bias voltage of sufficient magnitude

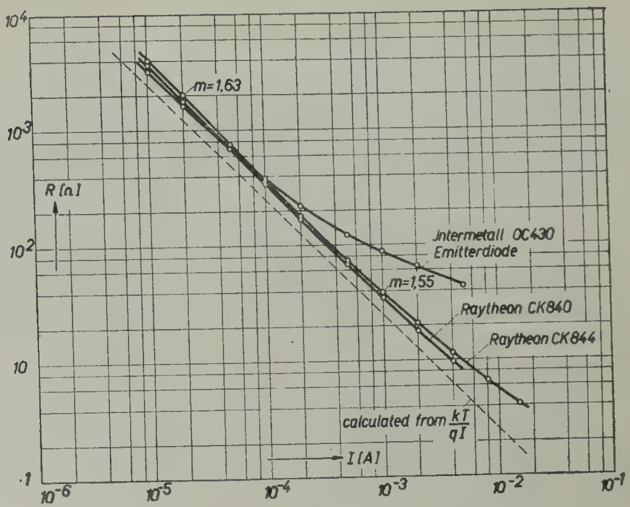

Fig. 2—Reciprocal value R of the real part of the diode admittance Y (vertical in ohms) vs diode current in amperes (horizontal) for three different silicon diodes. Some calculated values of m have been inserted [see (8)]. The broken curve pertains to pure diffusion.

$$\text{Re}\,(Y) \approx \frac{qI_D}{kT} + \frac{qI_R}{2kT} = \frac{qI}{mkT}. \tag{8}$$

By (8), a multiplicator m has been defined, which is obviously dependent on the ration I_D/I_R. But, as will be shown experimentally, m is approximately constant within a limited range of the current I in some practical cases.

In Fig. 1, some experimental curves are shown, from which it is seen that m is between 1 and 2. In Fig. 2 some experimental values of $R = 1/\text{Re}\,(Y)$ are shown. Except in one case (OC 430), the values of m are fairly constant within considerable ranges of the current I. The one case referred to obviously contains a series resistance, which affects the R-curve already from 0.1 ma upwards. Furthermore, high-level current injection starts at about 0.2 ma with this particular diode. With the diodes CK 840 and CK 844 of Fig. 2, series resistances are small and high-level current injection starts above 1 ma.

New Theoretical Expressions for Shot Noise in Silicon P-N Junction Diodes

We shall deal here with shot noise only, as distinct from flicker noise. The latter starts at zero frequency and is manifest up to some upper frequency, dependent on the purity of the material in question. With pure material, the upper frequency may be less than 1 kc, whereas it may be more than 1 mc with impure material. The intensity of flicker noise for a fixed frequency interval, Δf, is approximately proportional to the inverse mean frequency f of this interval, Δf.

A formula for the mean-square shot-noise current of a p-n junction diode, in which carrier motion is exclusively due to diffusion, has been derived

$$\overline{i_{nd}^2} = [4kT\,\text{Re}\,(Y_D) - 2qI_D]\Delta f. \tag{9}$$

This derivation was first published, using special assumptions, by van der Ziel,[9] and then was shown to be of general validity by Guggenbuehl and Strutt.[1,10]

Inserting the value of (3) for Re (Y_D) yields, in the case of a forward bias voltage (positive from p to n) of sufficient magnitude ($qV>3kT$)

$$\overline{i_{nd}^2} = 4qI_D\Delta f - 2qI_D\Delta f = 2qI_D\Delta f. \qquad (10)$$

These expressions (9) and (10) yield the mean-square value of a noise current i_{nd} due to a current generator, whereby the current generator is connected in parallel to the diode terminals.

It is, of course, also possible to describe the diode's shot noise by means of a noise voltage generator, connected in series to the diode. If Z_D is the impedance of the diode, then

$$\overline{v_{nd}^2} = 4kT \text{ Re } (Z_D)\Delta f - 2qI_D \left| Z_D \right|^2\Delta f. \qquad (11)$$

The derivation of the expression (11) may be given, along similar lines, as the derivation[1] of (9).

Tentatively, one might assume the deviation of expressions (10) and (11) from experimental results with silicon diodes to be due to a series resistance incorporated in the diode outside the space-charge layer. Assuming the series impedance to be Z_s, and the diode current I, with a diode impedance (of the space-charge layer) $1/Y$, we obtain, instead of (11)

$$\overline{v_{nd}^2} = 4kT \text{ Re}\left(\frac{1}{Y} + Z_s\right)\Delta f - 2qI\left|\frac{1}{Y}\right|^2\Delta f. \qquad (12)$$

Instead of a noise current generator, or of a noise voltage generator, the noise is also often described by the "equivalent noise resistance" R_{eq}. This is a resistance, connected by definition with the noise generator voltage v_{nd} by

$$\overline{v_{nd}^2} = 4kTR_{eq}\Delta f. \qquad (13)$$

We shall now relate this equivalent noise resistance to the differential admittance of the diode.

In the case of diffusion without series impedance, *i.e.*, according to (9) and (2) we obtain

$$\frac{R_{eq}}{R_D} = \frac{1}{2}\left(1 + \frac{I_o}{I_D + I_o}\right) \qquad (14)$$

if $V=0$ [see (1)], we have $I_D=0$, and, hence, $R_{eq}=R_D$. On the other hand, if $qV>3kT$, we have (forward bias) $I_D \gg I_0$, and, hence, $R_{eq}=\frac{1}{2}R_D$. This important relationship (14) was sufficiently confirmed experimentally with germanium junction diodes.[11]

If we have diffusion with series resistance $\text{Re}(Z_s)=R_s$, we obtain in the case of a forward bias of sufficient magnitude ($qV>3kT$)

$$R_{eq} = \tfrac{1}{2}R_D + R_s. \qquad (15)$$

Now we consider the case of diffusion plus recombination-generation. It has been shown that recombination-generation is predominantly located *at the center plane* of the depletion space-charge layer, if the mean lifetimes of carriers in the semiconductors adjacent to this layer are not too different.[4] The maximum of the recombination-generation rate U is exactly located at the center plane of the depletion layer, if the lifetimes, the mobilities, and the carrier densities are equal on opposite sides of the junction (symmetrical diode). In the case of an unsymmetrical diode the maximum of U is not appreciably displaced out of the center plane (see Appendix formula for distance d). We shall, for simplicity, assume generation-recombination to take place at the exact center plane of the depletion layer. The motion of carriers in this layer is assumed to be random without intercorrelation. A carrier is either captured (recombination) or generated at the center plane. In the first case a carrier moves from one of the boundary planes of the depletion layer to the center plane; in the second case it drifts from the center plane to either of the boundary planes. Thereby, each carrier which is generated or recombines in the depletion layer only goes through exactly one-half of the total thickness of this layer. In the diode circuit the effect is the same as if a carrier of charge $q/2$ were to go through the total depletion layer. According to W. Schottky's diode noise equation we obtain for the mean-square noise current $\overline{i_{np}^2}$ due to recombination-generation of holes, and for the mean-square noise current $\overline{i_{nn}^2}$ due to recombination-generation of electrons

$$\overline{i_{np}^2} = \overline{i_{nn}^2} = 2\,\frac{q}{2}\,\frac{I_R}{2}\,\Delta f.$$

Here, it is assumed that one-half of the total recombination current I_R is due to holes, and one-half to electrons. The total mean-square noise current $\overline{i_{nr}^2}$ due to recombination-generation is, hence

$$\overline{i_{nr}^2} = \overline{i_{np}^2} + \overline{i_{nn}^2} = qI_R\Delta f. \qquad (16)$$

We may apply (16) in order to obtain the equivalent of (9) in the case of recombination-generation shot noise. The result is[12]

$$\overline{i_{nr}^2} = 4kT \text{ Re } (Y_R)\Delta f - qI_R\Delta f. \qquad (17)$$

If we add the mean-square shot-noise current due to diffusion to the above expressions (16) and (17), bearing in mind that the two shot noise currents are incorrelated, we obtain

[9] A. van der Ziel, "Shot noise in junction diodes and transistors," Proc. IRE, vol. 43, pp. 1639–1646; November, 1955, and vol. 45, p. 1011; July, 1957.

[10] W. Guggenbuehl and M. J. O. Strutt, "Theorie des Hochfrequenzrauschens von Transistoren bei kleinen Stromdichten," Nachrtech. Fachberichte, Beiheft der NTZ, vol. 5, pp. 30–33; September, 1956.

[11] Guggenbuehl and Strutt, *op. cit.*, Fig. 18, footnote 1.

[12] *Ibid.*, pp. 839–841.

$$\overline{i_n^2} = \overline{i_{nd}^2} + \overline{i_{nr}^2} = 4kT[\text{Re}\,(Y_R) + \text{Re}\,(Y_D)]\Delta f - 2q\left(I_D + \frac{I_R}{2}\right)\Delta f. \tag{18}$$

Here again, we apply (8) for the multiplication factor m and obtain

$$\overline{i_n^2} = 4kT\,\text{Re}\,(Y) - 2qI\frac{\Delta f}{m}, \tag{19}$$

where I is the total diode current and Y the total diode admittance. In the low-frequency region, we have, in the case of forward bias of sufficent magnitude ($qV > 3kT$)

$$\text{Re}\,(Y) \approx \frac{qI}{mkT}.$$

Hence, (19) in this case reduces to

$$\overline{i_n^2} = 2q\frac{I}{m}\Delta f. \tag{20}$$

This (20) may again be reverted to an equivalent noise resistance R_{eq}. We then obtain, if Re $(Y) = 1/R$

$$\frac{R_{eq}}{R} \approx \frac{1}{2}, \tag{21}$$

in the case of a forward bias of sufficient magnitude ($qV > 3kT$). It is noteworthy that the relation (21) in the case of diffusion plus recombination noise is quite similar to (14) for a forward bias of sufficient magnitude ($qV > 3kT$).

If an appreciable series resistance R_s is present in the diode, (21) must be altered slightly, correspondingly to (15). In this case, $R + R_s$ denotes the total diode resistance, and we have

$$R_{eq} = \tfrac{1}{2}R + R_s. \tag{22}$$

This equation may be extended so as to obtain the equivalent of (12) in the case that recombination-generation noise is included. We obtain

$$\overline{v_n^2} = 4kT\,\text{Re}\left(\frac{1}{Y} + Z_s\right)\Delta f - 2q\frac{I}{m}|Z|^2\Delta f. \tag{12a}$$

If we take into account that recombination-generation of carriers is not concentrated at the center plane of the depletion layer, but has a certain distribution over the depletion layer, (16) is somewhat altered (see Appendix). However, the theoretical deviations from (16) are small. Likewise, if the recombination-generation is assumed to be concentrated at a plane, which, however, is somewhat displaced with regard to the center of the depletion layer, (16) is somewhat altered (see Appendix). In this case, also, the theoretical deviation from (16) is small.

New Theoretical Expressions for Shot Noise in Silicon Junction Transistors

We consider junction transistors in the three possible connections: grounded base, grounded emitter, and grounded collector. First, the intrinsic transistor, consisting of emitter and of collector diode, biased properly, is treated. The admittance of the emitter diode is denoted by Y_{11}. Then, (19), applied to this diode, yields

$$|\overline{i_{ne}^2}| = 4kT\,\text{Re}\,(Y_{11})\Delta f - 2q\frac{I_E}{m_E}\Delta f, \tag{23}$$

where I_E is the emitter dc current, and m_E is the multiplication factor [see (8)] of the emitter diode.

The collector current is composed of two parts: first, the saturation current I_{co} caused by generation of carriers at the edges of the depletion layer and in the depletion layer, second, the injected current from the emitter into the base layer minus recombination current in this layer. The first part is relatively so small in silicon transistors that it may be neglected as to its noise contribution. Hence, the collector diode gives rise to a noise-current generator

$$|\overline{i_{nc}^2}| = 2qI_c\Delta f. \tag{24}$$

The correlation between the two noise currents i_{ne} and i_{nc} is given by the equation

$$\overline{i_{ne}^* i_{nc}} = 2kTY_{21}\Delta f. \tag{24a}$$

The deviation may be carried out in exactly the same manner as has been done.[13] In (24a) the term $-2kTY_{12}^*\Delta f$ has been neglected at the right side. Applying (23), (24) and (24a), according to calculations along exactly the same line as published previously,[1,10] we obtain

$$4kTR_oF = 2qI_c\left|\frac{\frac{1}{Y_{11}} + R_b + R_o}{\alpha_{fb}}\right|^2 - 2q\frac{I_E}{m_E}|R_b + R_o|^2. \tag{25}$$

Here, F is the noise figure for the grounded base and for the grounded emitter connection, R_o is the source resistance (assumed to be at the same absolute temperature T, as the transistor), R_b the extrinsic base resistance, and α_{fb} the ac current amplification factor in the grounded base connection. The conditions, involved in (24), (24a) and (25) are similar to those stated for pure diffusion transistors[1,10]

$$|Y_{11}| \gg |Y_{12}|, \quad |Y_{21}| \gg |Y_{22}|, \quad |Y_{22}R_b| \ll 1. \tag{26}$$

Here, Y_{11}, Y_{12}, Y_{21}, and Y_{22} are the admittances of the intrinsic transistor, as usual.

For the grounded collector connection, we may again

[13] *Ibid.*, p. 843.

apply the procedure, published previously[1,10] bearing in mind (23), (24), and (24a). We then obtain

$$4kTR_oF = 4kT(R_o + R_b) + 2qI_c(R_o + R_b)^2 - |\alpha_{fb}|^2 2q \times \frac{I_E}{m_E}\left|R_o + R_b - \frac{1}{Y_{21}}\right|^2 - 4kT|\alpha_{fb}|^2 \operatorname{Re}\left(\frac{R_o}{\alpha_{fb}}\right) - 4kT|\alpha_{fb}|^2 R_b \operatorname{Re}\left(\frac{1}{\alpha_{fb}}\right) + \frac{4kT}{|Y_{11}|^2}\operatorname{Re}(Y_{11}). \quad (27)$$

The conditions for (27) are again (26).

Experimental Confirmations of Theoretical Noise Expressions for Silicon Junction Diodes and Transistors

In order to obtain the above experimental confirmation, a measuring set was constructed, with which the noise of junction diodes, biased in the forward direction, may be determined. The equivalent noise resistance of such objects is often below the input noise resistance of low-noise tube circuits. Hence, a suitable transformer must be applied between the diode to be measured and the input tube circuit of the measuring set.[14]

The measuring set is shown schematically in Fig. 3.

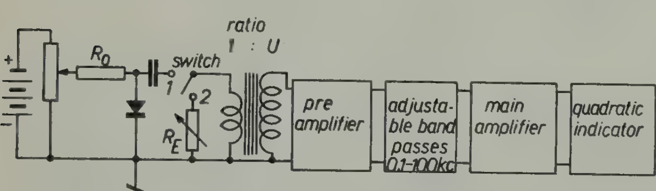

Fig. 3—Measuring setup for the determination of diode noise. For a description, see text.

If the switch at the input is connected to contact 1, an indication is obtained at the output indicator, due to diode noise, amplifier noise, and noise of the resistive parts of the transformer impedances. Then the switch is connected to contact 2, and the purely ohmic resistance R_E is adjusted until the same output indication is obtained. The value of R_E is determined and is then equal to the equivalent noise resistance of the diode under consideration. This determination of the diode's equivalent noise resistance R_{eq} may be carried out with sufficient accuracy if

$$R_{eq}u^2 \geqq R_{eq\ ampl} + R_{o2} + R_{o1}u^2$$

Here, u is the transformer ratio, $R_{eq\ ampl}$ is the equivalent input noise resistance of the preamplifier, R_{o2} is the secondary, and R_{o1} the primary effective resistance of the transformer windings. The upper frequency of the set is connected with the leakage inductance of the transformer. It has been possible, by this method, to determine equivalent noise resistances of less than one ohm with reasonable accuracy.

Some measured values of R_{eq} for silicon junction diodes are shown in Figs. 4, 5 and 6. As is seen from Fig. 4, the value of R_{eq} coincides satisfactorily with the value $\frac{1}{2}R$, *i.e.* with one-half of the diode's resistance, according to (21). On the other hand, the measured values of R_{eq} are distinctly above those, calculated from (10). In Fig. 5, measured values of R_{eq} as dependent on frequency at different dc values of the diode current I are shown. The values of R_{eq} tend to drop above about 20 kc. This is due to the frequency dependence of the diode's impedance $1/Y$. It is seen from Fig. 5, that R is still independent of frequency at the frequency, at which the values of Figs. 4 and 6 have been measured. At higher currents, a deviation of measured values of R_{eq} from $\frac{1}{2}R$ is shown in Fig. 4 and Fig. 6. This is due to high-level current injection, which invalidates the basis of our calculation.

The silicon diode Raytheon type 1 N 301 is manufactured according to the "bonded" procedure. It has a series resistance of about 26 ohms. The real part of its impedance $1/Y$ is independent of frequency up to above 70 kc. The curve a of Fig. 7 was calculated from (22), taking into account the said series resistance. This curve a coincides approximately with the measured values of R_{eq}, except at higher current values. This is due to high-level current injection, which starts below 1 ma with this diode. As a proof hereof, Fig. 8 shows measured values of the susceptance of this diode as dependent on dc diode current I. The susceptance goes negative at values of I above 1 ma, which is a sure sign for high-level current injection.[8] Of course, high-level injection starts before the susceptance becomes negative. If we use (12) for the calculation of R_{eq}, curve b of Fig. 7 is obtained, which shows definitely worse agreement with measured values of R_{eq} than curve a.

In order to apply (25) to calculate the noise figure of transistors, the value of m_E must be known. The experimental determination of m_E is not straightforward. One method makes use of measurements of the differential emitter diode admittance. According to (8), the value of m_E may then be deduced. However, this method yields accurate values of m_E only for small current values. As soon as high-level current injection starts, (8) is no longer satisfied. Besides, an ohmic series resistance, which cannot be exactly determined, will be of increasing importance with increasing forward current. As an example, reference is made to Fig. 2, in which one curve pertains to the emitter diode of a silicon transistor type OC 430. Here, deviations from (8) already occur above a dc emitter current of 0.1 ma.

Another method for the determination of m_E could be based approximately on measurements of the dc current amplification factor α_{FB}. The emitter current of silicon transistors consists of two parts, one due to dif-

[14] W. Nonnenmacher, "Rauschspannungsmessungen an niederohmigen Bauelementen mit Hilfe eines Röhrenverstärkers mit vorgeschaltetem Uebertrager," Nachrtech. *Z.* vol. 11, pp. 559–563; November, 1957.

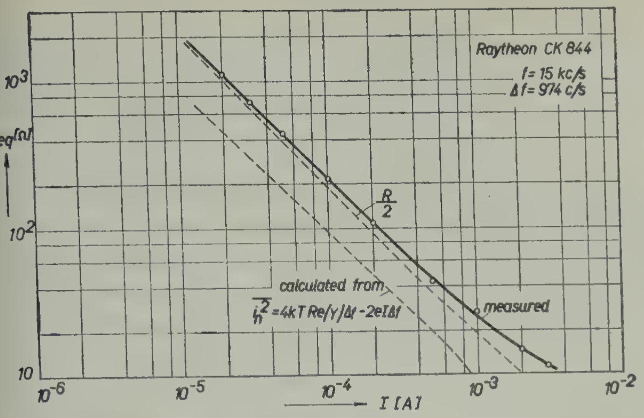

Fig. 4—Equivalent noise resistance R_{eq} in ohms (vertical) as measured (full curve) as dependent on diode current in amperes (horizontal). The broken curve marked $\frac{1}{2}R$ is calculated by taking one-half of the reciprocal value R of the real part of the measured diode differential admittance. The other broken curve is calculated from (10).

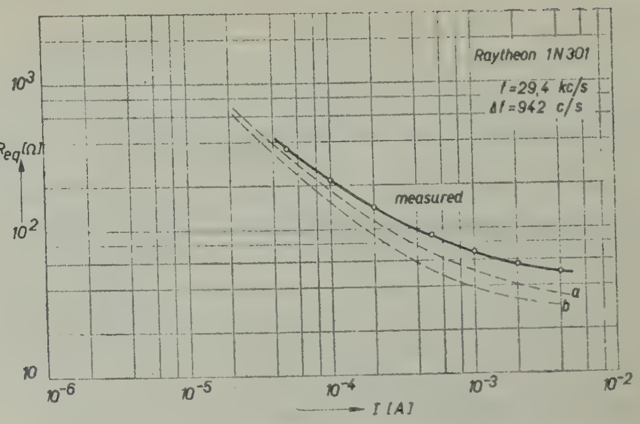

Fig. 7—Similar to Fig. 4, but for again another type of diode. Full curve is measured values of R_{eq}. Broken curve marked a is calculated from (22). Broken curve marked b is calculated from (12).

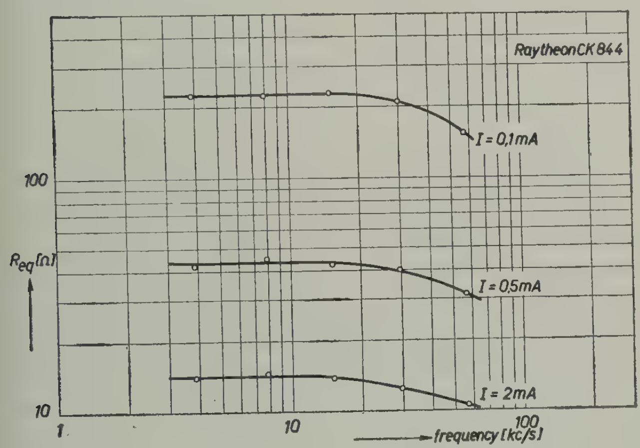

Fig. 5—Measured equivalent noise resistance R_{eq} as dependent on frequency at three values of diode current.

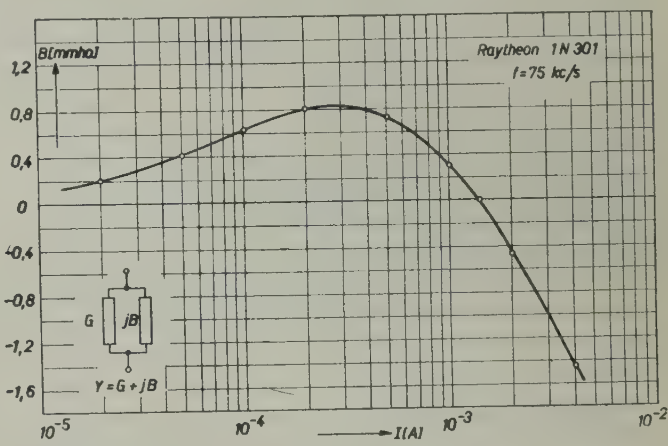

Fig. 8—Differential susceptance in mmhos (vertical) as dependent on diode current (horizontal), of a silicon diode.

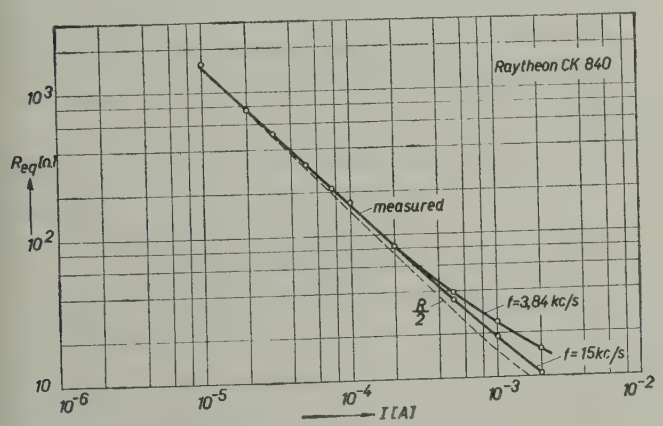

Fig. 6—Similar to Fig. 4, but for another type of diode. Measured values of R_{eq} at two different frequencies are full curves. The deviation between the curves $f=3.84$ kc and f=15 kc is caused by flicker noise. Broken curve is calculated from $\frac{1}{2}R$.

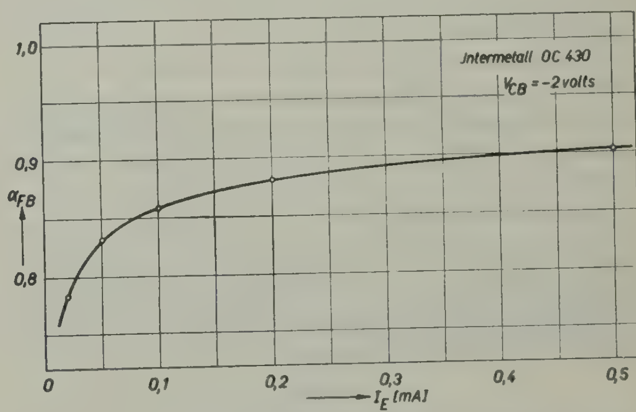

Fig. 9—DC current amplification factor α_{FB} of a silicon transistor, as dependent on emitter current.

fusion and the other due to recombination-generation. As the latter part consists of majority carriers on the n-side as well as on the p-side of the depletion layer, it offers no contribution to the minority current injected into the base layer. If the emitter current diminishes, the recombination-generation current becomes more important in comparison to the diffusion current. Hence, the value of α_{FB} is correspondingly lower. This is confirmed by measurements, shown in Fig. 9. In order to evaluate the recombination current part of the emitter current, and hence, m_E from α_{FB}, the emitter efficiency, the surface recombination and the volume recombination of the base layer should be known, which is hardly

ever true. Thus, this method is not applicable.

If α_{FB} is near unity, recombination in the emitter depletion layer must be slight. Measured values of the noise figure F in grounded base connection of an alloyed silicon transistor type OC 470 are shown in Fig. 10 as being dependent on frequency. In the calculated curve, recombination *was neglected*, *i.e.*, $m_E=1$ in (25). The discrepancy between the curves of Fig. 10 is not very marked. It is partly due to flicker noise inherent in the measured curve at lower frequencies. At higher frequencies the measured curve rises less steeply than the calculated curve. This is due to parallel capacity to the base resistance, which occurs in this transistor.

In Fig. 11, which is similar to Fig. 10 but for a different type of transistor, having a lower value of α_{FB}, recombination must be taken into account in order to obtain calculated curves, which are reasonably near to measured values. At high frequencies the rise of the measured curve is less steep than that of calculated curves, due to the capacity parallel to R_b, as mentioned above.

In both Figs. 10 and 11, measured values are above calculated values, taking recombination into account. This may be ascribed to high-level emitter current injection. As is shown in Fig. 12, the curve, calculated with recombination, fits the experimental values well at very low currents and deviates at higher currents. The curve, calculated without recombination, fits experimental values badly.

In cases of high-level current injection, the theory set forth here, which pertains to low current densities (no space charge caused by carriers), is only in part agreement with measured values. A better agreement could be expected from an extension of the theory to high level current injection.

A conjecture has been made previously[1] on the influence of high-level injection on noise. This conjecture may be extended here to include the case of recombination-generation noise. Referring to the above expression (19), the space charge in the semiconductor portions adjacent to the depletion layer, created by high-level injection, will probably tend to weaken the contribution embodied in the second term of (19). Hence, the total mean-square noise current will rise. This seems to be confirmed by the Figs. 4, 6, 7, 10, and 11.

Appendix

Calculation of Noise Due to Recombination-Generation

Assuming a linear potential variation across the junction, the rate U of recombination-generation in the depletion layer is approximately given by:

$$U(\xi) \approx \frac{n_i \exp\left(\frac{qV}{2kT}\right)}{\tau_{po} \exp Q + \tau_{no} \exp(-Q)},$$

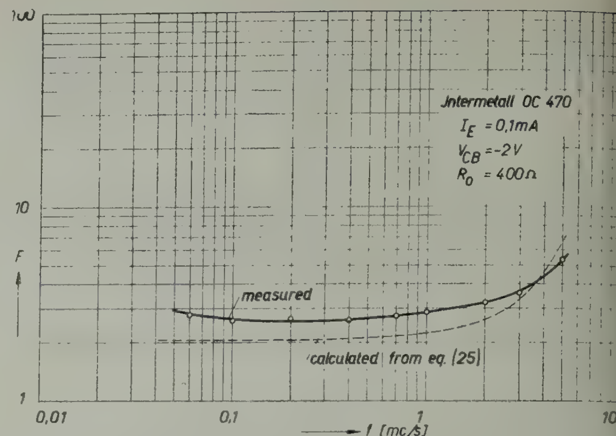

Fig. 10—Noise figure F (vertical) of a silicon transistor at fixed values, as indicated, of emitter current I_E, voltage V_{CB} and source resistance R_o, measured in grounded base connection, as dependent on frequency (horizontal). The value of α_{fbo} of α_{fb} at low frequency is 0.970. The broken curve is calculated from (25).

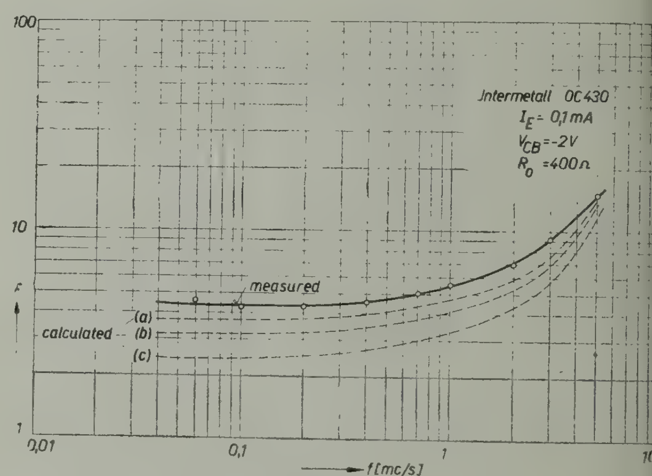

Fig. 11—Similar to Fig. 10, but for another type of silicon transistor. Here the value of α_{fbo} is 0.885. The calculated broken curves are a) from (25) inserting the most probable value of m_E, b) from (25) at $m_E=1$, but inserting the most probable value of Y_{11} and c) at $m_E=1$ and with a value of Y_{11} ignoring recombination.

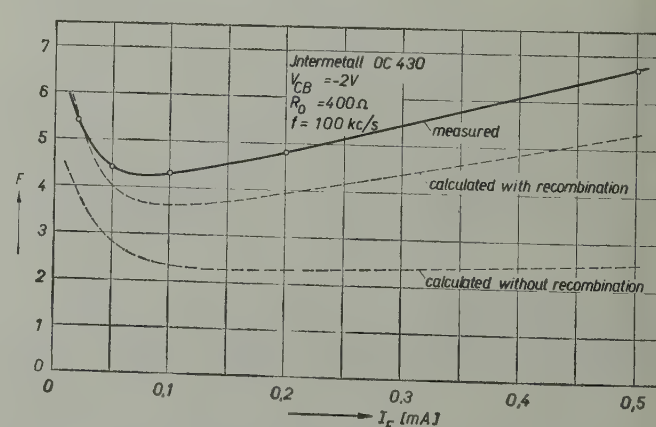

Fig. 12—Noise figure F (vertical) of the same silicon transistor of Fig. 11, as dependent on emitter current at fixed values, as indicated, of V_{CB}, R_o and frequency. Broken curves calculated as indicated.

where

$$Q = \frac{q}{kT} \cdot \frac{V_D - V}{w} \xi.$$

V_D is the diffusion voltage, w is the width of the depletion layer, ξ a coordinate along the axis of the depletion layer, which is zero at the maximum of U, τ_{po} is the lifetime of holes injected into n-type material of high dotation, τ_{no} is the lifetime of electrons injected into p-type material of high dotation, n_i is the density of electrons or holes in intrinsic material.

A picture of the depletion layer with densities of acceptors, donors, holes, and electrons is shown in Fig. 13. The maximum of U, if the lifetimes τ_{po} and τ_{no} are not too different, is at the place where $p(x) = n(x)$.[15] Taking account of the following relations for w_p and w_n, we obtain for the distance d,

$$w_n \approx \frac{w p_p}{n_n + p_p} \qquad w_p \approx \frac{w n_n}{n_n + p_p}$$

$$d = w_p + x_1 = \frac{w}{2}\left[1 + \frac{kT}{q(V_D - V)} \ln \frac{p_p}{n_n}\right]. \tag{28}$$

The second term within square brackets is usually small with respect to unity and, therefore, the maximum of U in unsymmetrical diodes is not considerably displaced out of the center of the depletion layer.

First we consider the influence on noise of the distribution over the depletion layer of the carrier-recombination-generation. To carry this out, we suppose the maximum of U to be at the center of the depletion layer (symmetrical diode). In a part of the depletion layer between x and $x+dx$ the recombination current contribution is

$$dI_R = qU(x)Adx,$$

where A is the cross-section area of the junction. This contribution dI_R may be separated into a part, due to electrons

$$dI_{Rn} = q \frac{\frac{w}{2} - x}{w} U(x)Adx,$$

and a part, due to holes

$$dI_{Rp} = q \frac{\frac{w}{2} + x}{w} U(x)Adx.$$

These parts each involve a noise current, the mean-square values of which are

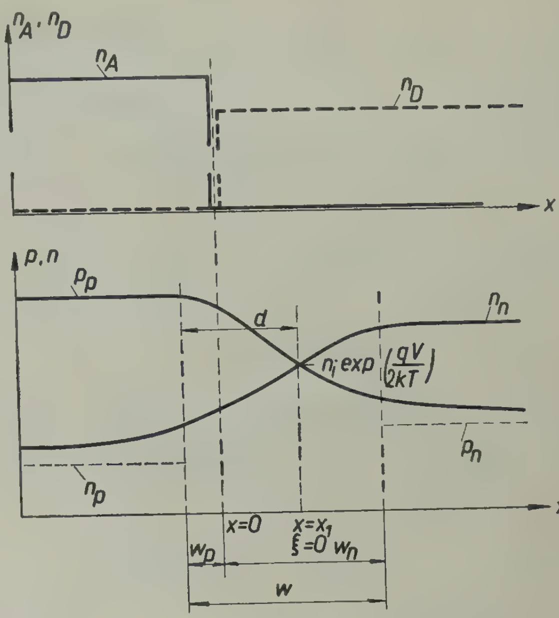

Fig. 13—The upper part shows the density n_A of acceptors, and the density n_D of donors in a cross-section of a p-n junction diode. The lower part shows the density n of electrons and the density p of holes in a p-n junction diode, where n_n is the density of electrons in n-material at thermal equilibrium, n_p the density of electrons in p-material at thermal equilibrium, p_n the density of holes in n-material at thermal equilibrium and p_p the density of holes in p-material at thermal equilibrium. The depletion layer is between $\xi = -x_1 - w_p$ and $\xi = w_n - x_1$.

$$d(\overline{i_{nn}^2}) = 2q \frac{\frac{w}{2} - x}{w} q \frac{\frac{w}{2} - x}{w} U(x)Adx\Delta f,$$

and

$$d(\overline{i_{np}^2}) = 2q \frac{\frac{w}{2} + x}{w} q \frac{\frac{w}{2} + x}{w} U(x)Adx\Delta f.$$

The total differential mean-square noise current due to recombination-generation is

$$d(\overline{i_{nr}^2}) = q^2 A U(x)\left[1 + \left(2 \frac{x}{w}\right)^2\right] dx\Delta f.$$

This should be integrated along the depletion layer (see Fig. 13) from $x = -(w/2)$ to $x = +(w/2)$, assuming that $w_p = w_n$ in Fig. 13, *i.e.*, that $n_A = n_D$. By symmetry, and owing to the rapid decrease of $U(x)$ on both sides of $x = 0$, we obtain approximately

$$\overline{i_{nr}^2} = 2q^2 A\Delta f \int_0^\infty U(x)\left[1 + \left(2 \frac{x}{w}\right)^2\right] dx$$

or

$$\overline{i_{nr}^2} = qI_R\Delta f\left[1 + 8\left(\frac{kT}{q(V_D - V)}\right)^2\right]. \tag{29}$$

The second term within square brackets is usually small, compared with the first one, and hence, (29) yields approximately (16).

[15] Sah, Noyce, and Shockley, *loc. cit.*, pp. 1230–1231.

We now assume that recombination-generation is concentrated at $\xi=0$, but that this place is not the center of the depletion layer (unsymmetrical diode) (see Fig. 13). The electron contribution to the total recombination current is (see Fig. 13):

$$I_{Rn} = I_R \frac{w-d}{w}$$

and the hole contribution is

$$I_{Rp} = I_R \frac{d}{w}.$$

The path of the electrons being $w-d$ and that of the holes d, the corresponding mean square noise currents are

$$\overline{i_{nn}^2} = 2q \frac{w-d}{w} I_R \frac{w-d}{w} \Delta f,$$

$$\overline{i_{np}^2} = 2q \frac{d}{w} I_R \frac{d}{w} \Delta f.$$

Inserting (28) for the distance d, we obtain

$$\overline{i_{nr}^2} = qI_R(1 + c^2)\Delta f,$$

where

$$c = \frac{kT}{q(V_D - V)} \cdot \ln \frac{p_p}{n_n}.$$

Again, the second term within brackets is small with respect to the first one and, hence, we obtain approximately (16).

List of Symbols

τ_{po}—lifetime for holes injected into n-type material of high dotation.
τ_{no}—lifetime for electrons injected into p-type material of high dotation.
τ_o—lifetime for electrons and holes when $\tau_{po}=\tau_{no}$.
U—recombination rate.
V_D—diffusion voltage in a p-n junction.
m—multiplication factor to describe the empirical characteristics.
m_E—multiplication factor of the emitter junction in transistors.
w—width of the space-charge layer.
w_n—n-material portion of the space-charge layer.
w_p—p-material portion of the space-charge layer.
I_D—diffusion current in a p-n junction.
I_R—recombination current in a p-n junction.
I—total current in a p-n junction.
Y_D—diffusion differential admittance.
Y_R—recombination differential admittance.
Y—Y_D+Y_R=total differential admittance of a p-n junction.
Z_D—$1/Y_D$=diffusion differential impedance.
Z_S—series impedance of the p-n junction.
R_D—Re (Z_D) diffusion differential resistance.
R—1/Re (Y) at low frequencies=total differential resistance.
R_s—series resistance.
i_{np}—noise current caused by holes.
i_{nn}—noise current caused by electrons.
i_{nr}—recombination noise current.
i_{nd}—diffusion noise current.
i_{ne}—noise current of the emitter diode.
i_{nc}—noise current of the collector diode.
v_n—noise voltage.
v_{nd}—diffusion noise voltage.
α_{fb}—ac current amplification factor in grounded base connection.
α_{fb}—dc current amplification factor in grounded base connection.

Acknowledgment

The authors wish to express their sincere thanks to G. Spescha who has contributed to the experimental setup of Fig. 3.

A Constant-Temperature-Operation Hot-Wire Anemometer*

J. M. L. JANSSEN†, SENIOR MEMBER, IRE, L. ENSING†, MEMBER, IRE AND J. B. VAN ERP†

Summary—The behavior of a hot-wire anemometer is analyzed using signal flow diagrams, especially for the case of a constant-temperature-operation anemometer, comprising a dc bridge and a dc differential amplifier.

The influence of static bridge unbalance and of the amplifier's gain for in-phase input signals on the over-all dynamic characteristics is analyzed, and requirements for the amplifier are derived. A chopper-stabilized amplifier is employed, which automatically maintains static bridge balance under all operating conditions.

The signal-to-noise ratio for comparable instruments having constant-current operation and constant-temperature operation is shown to be equal.

Results of measurements are reported; they are in fair agreement with the developed theory.

A nonlinear circuit is described, which delivers a voltage directly proportional to the instantaneous air velocity.

List of Symbols

M = time constant of hot wire.
M', M'' = main time constant of over-all control system.
ω = angular frequency.
C = heat capacitance of hot wire.
λ = quantity as defined by (18).
a, b = constants occurring in King's equation.
Φ, Ψ = quantities as defined by (64) and (65).
R = hot-wire resistance at operating temperature.
R_a = hot-wire resistance at air temperature.
R_1, R_2, R_3 = resistances of the branches of the bridge (see Fig. 2).
T = hot-wire temperature.
T_a = temperature of the air.
I = total bridge current.
I_a = bridge current supplied by amplifier.
I_e = bridge current supplied by external source.
E = voltage across hot wire.
E_1 = voltage across bridge resistance R_1.
E_b = bridge unbalance voltage.
E_d = equivalent drift at input of amplifier.
E_i = in-phase component of the amplifier's input voltage.
P = heat flow to wire due to electrical heating ($=I^2R$).
H = heat flow from wire due to cooling.
W = resulting heat flow to wire ($=P-H$).
S, S', S_0 = loop gain.
S_d, S_d' = dynamic loop gain.
q = quantity as defined by (40).
f = rejection factor of difference amplifier [see (38)].
Λ = relative static bridge unbalance [see (35)].
g_0 = static transconductance of controlling amplifier.
K_1 = quantity as defined by (4).
τ = per-stage time constant of amplifier.
cco = abbreviation for "constant-current operation."
cto = abbreviation for "constant-temperature operation."
$r, t, i, i_a, i_e, e, e_1, e_b, p, h, w$, etc., are variations of $R, T, I, I_a, I_e, E, E_1, E_b, P, H, W$, etc., respectively.

I. Introduction

IN the investigation of turbulent air flow use is often made of hot-wire anemometers. These instruments are based on the cooling effect of the air stream on a heated wire. The wire is heated by an electrical current and placed in the air flow; variations in the velocity of the air flow, due to turbulence, cause the resistance of the wire to vary. The variations in the wire's resistance give rise, in turn, to variations in the voltage across the wire, which after being amplified give an image of the air velocity variations. Thus the characteristic properties of the flow can be determined.

In order that the variations in the wire's resistance may follow the variations in air velocity as closely as possible, it is necessary to keep the heat capacitance of the hot wire per unit length as small as possible. For this reason very thin wires (diameter 3 μ) are used. These wires are part of a probe, which is introduced into the turbulent flow. To fix the wire to the probe a special spot-welding technique has been developed [1].

In view of the small dimensions of the eddies in a turbulent flow the length of the hot wire has to be limited. Otherwise the effects of several eddies on the wire resistance are bound to neutralize each other to a certain extent. A length of about 1 mm would be appropriate. If a tungsten wire with a diameter of 3 μ is used, the resistance at room temperature will then be about 10 ohms.

Two different methods [2]–[6] can be distinguished in hot-wire anemometry: 1) constant-current operation (referred to in this paper as cco), and 2) constant-temperature operation (here referred to as cto).

With cco (see Fig. 1) the ratio of voltage variations across the hot wire to air velocity variations drops off with frequency according to $1/\sqrt{1+(\omega M)^2}$, M being the time constant of the wire and ω being the angular fre-

* Original manuscript received by the IRE, September 19, 1958.
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quency. In order to compensate for this linear distortion, the signal has to be passed through a filter with an inverse frequency characteristic. M, however, is not a constant for a particular wire but varies with wire temperature and air velocity. For this reason the filter characteristics must be readjusted if the operating conditions are changed. This is very time-consuming and tedious work. Compensation becomes increasingly inaccurate for large air velocity variations because M can then no longer be considered a constant during a single cycle.

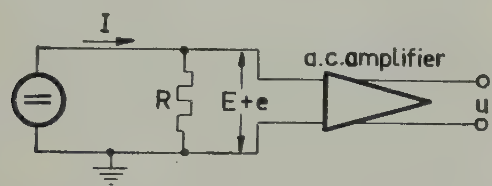


Fig. 1—Hot-wire anemometer with constant-current operation.

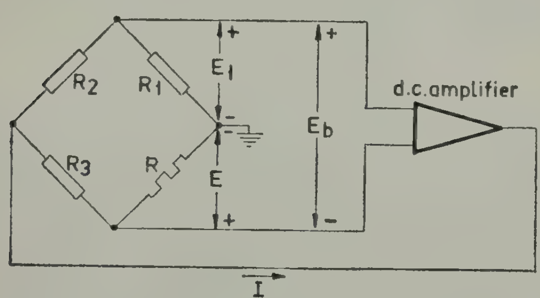


Fig. 2—Hot-wire anemometer with constant-temperature operation.

The necessity to study large velocity fluctuations containing frequencies of several tens of kilocycles per second has led to the development of cto anemometers. The wire is incorporated into a feedback system, *e.g.*, by means of a bridge circuit and dc amplifier (see Fig. 2). In this way the main time constant of the instrument becomes $M/1+S$, where M is the time constant of the hot wire and S is the loop gain. The fact that the controlling amplifier introduces an additional phase lag at high frequencies gives rise to stability problems.

Eli Ossofsky [6] derived the conditions that have to be complied with by the dc amplifier of Fig. 2 in order to avoid parasitic high-frequency oscillations. He starts from the assumption that the bridge is balanced for the adjusted mean values of air velocity and heating current. In our experience, it was found to be impossible to maintain the bridge in balance during measurement. Moreover, it was found that a slight bridge unbalance often gave rise to self-oscillation of the feedback system. This led to an investigation of the influence of the static bridge unbalance on the dynamic characteristics. From this study, the requirements regarding the static gain and the zero stability of the control amplifier were derived.

Based on this study, a cto hot-wire anemometer was constructed, which has the novel features of drift-stabilization and linearizing.

A chopper stabilizing circuit is provided in the controlling amplifier in order to maintain bridge balance under all operating conditions.

The relation between the air velocity and the hot-wire current with cto is nonlinear. At small amplitudes, therefore, the ratio between bridge current variations and air velocity variations changes with the static value of the air velocity, and at large amplitudes nonlinear distortion occurs. In the constructed instrument, this nonlinearity is compensated by a nonlinear circuit with an inverse characteristic which derives a voltage directly proportional to the air velocity from the bridge current.

II. Frequency Response of the Hot Wire

In this section the response of the hot-wire temperature to variations in air velocity is discussed. Since the temperature of the wire determines its resistance, this section covers at the same time the response of the hot-wire resistance to variations in air velocity. Knowledge of this resistance response enables one to find the response of the total electric measuring circuit in which the hot wire is incorporated.

Fig. 3 shows a signal flow graph [7] of the hot wire. The difference between the heat flows to and from the wire "charges" or "discharges" its heat capacitance C, giving rise to a temperature variation t. The associated transfer function is $1/j\omega C$.

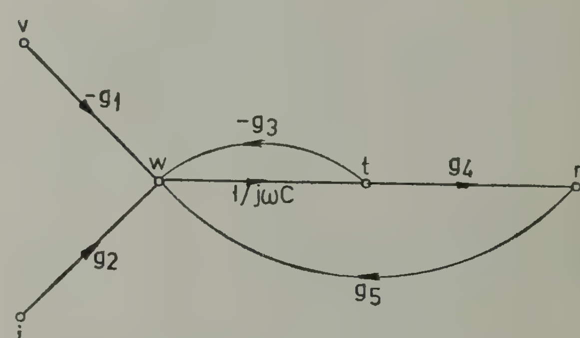


Fig. 3—Signal flow graph of hot wire.

There are four heat flows which in the analysis can be considered separately, although physically they are interdependent as shown in Fig. 3. They are: 1) the cooling due to an increase v in air velocity, 2) the heating due to an increase i in the current through the wire, 3) the cooling due to an increase t in temperature, and 4) the heating due to an increase r in wire resistance.

There are no time effects in any of the above-mentioned heat flows. Thus the associated transfer functions indicated in Fig. 3 are all constants. The same is true for the transfer function g_4 relating resistance to temperature variations.

The signal flow graph shows a negative feedback $-g_3$ and a positive feedback $g_4 \cdot g_5$. The stability condition for a constant current anemometer ($i=0$) is $g_4 \cdot g_5 < g_3$,

which condition is always satisfied in practice [see discussion of (22)].

The transfer function from v to r is

$$\frac{r}{v} = -g_1 \cdot \frac{r}{w} = -g_1 \cdot \frac{g_4 \cdot M/C}{1 + j\omega M} \tag{1}$$

with

$$M = \frac{C}{g_3 - g_4 \cdot g_5}. \tag{2}$$

The above result is in agreement with results found by Betchov [8].

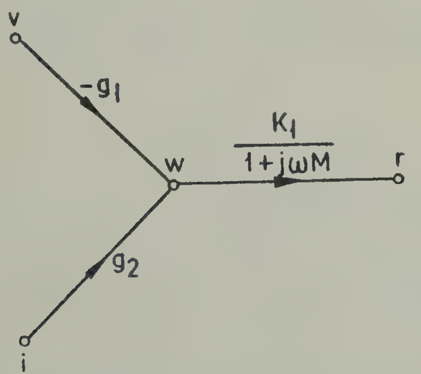
In constant-temperature operation another transfer function is equally of interest, namely

$$\frac{r}{i} = g_2 \cdot \frac{r}{w}. \tag{3}$$

It has the factor

$$\frac{r}{w} = \frac{g_4 \cdot M/C}{1 + j\omega M} = \frac{K_1}{1 + j\omega M} \tag{4}$$

in common with the transfer function r/v, as is shown in the simplified signal flow graph of Fig. 4.



$$\frac{r}{w} = \frac{g_4 \cdot M/c}{1 + j\omega M} = \frac{k_1}{1 + j\omega M}$$

Fig. 4—Simplified signal flow graph of hot wire.

The constants g_1 and g_3 can be found from King's equation which describes the cooling of a small cylinder in an air flow [9]. The constants g_2 and g_5 follow directly from Joule's law, while the constant g_4 is determined by the temperature coefficient of the wire resistance.

King's equation reads:

$$H = (a + b\sqrt{V})(T - T_a) \tag{5}$$

in which

H = the heat flow from the wire.
V = the air velocity.
T = the wire temperature.
T_a = the air temperature.
a and b are constants, dependent on wire dimensions.

If P is the heat flow to the wire due to electrical heating, then the following equation holds under static conditions

$$H = P = I^2R \tag{6}$$

where

I = the current through the wire.
R = the resistance of the wire.

It follows from (5)

$$h = \frac{b \cdot (T - T_a)}{2\sqrt{V}} \cdot v + (a + b\sqrt{V})t. \tag{7}$$

This equation also holds when the temperature is a function of the place along the wire because all quantities in the right-hand part of (7) except t and T are constant along the wire. In that case T stands for the average temperature and t for the average temperature variation. Hence, in Fig. 3:

$$g_1 = \frac{b(T - T_a)}{2\sqrt{V}} \tag{8}$$

$$g_3 = a + b\sqrt{V}. \tag{9}$$

As

$$p = d(I^2R) = 2IRi + I^2r \tag{10}$$

it follows that

$$g_2 = 2IR = \frac{2P}{I} \tag{11}$$

$$g_5 = I^2. \tag{12}$$

Further

$$g_4 = \frac{dR}{dT} = \frac{r}{t} \tag{13}$$

at the values of R, T and V at which the operation takes place.

Substitution of the equations for g_1, g_2, g_3 and g_5 in those for the time constant M [see (2)] and in the transfer functions r/w [see (4)], r/v [see (1)] and r/i [see (3)], yields, if use is made of (6):

$$M = \frac{C(T - T_a)}{P(1 - \lambda)} \tag{14}$$

$$\frac{r}{w} = \frac{R}{P} \cdot \frac{\lambda}{1 - \lambda} \cdot \frac{1}{1 + j\omega M} \tag{15}$$

$$\frac{r}{v} = \frac{R}{2V} \cdot \frac{b\sqrt{V}}{a + b\sqrt{V}} \cdot \frac{\lambda}{1 - \lambda} \cdot \frac{1}{1 + j\omega M} \tag{16}$$

$$\frac{r}{i} = 2 \cdot \frac{R}{I} \cdot \frac{\lambda}{1 - \lambda} \cdot \frac{1}{1 + j\omega M} \tag{17}$$

with

$$\lambda = \frac{g_4 \cdot (T - T_a)}{R}. \tag{18}$$

Eqs. (15), (16) and (17) could be rewritten as transfer functions for the relative variations in r, w, v and i by transferring R, P, V and I to the left hand side of the equations. Then the basic difference between these equations would lie only in the fact that the factor $b\sqrt{V}/a+b\sqrt{V}$ is present in (16) and absent in the others. This factor accounts for the fact that only the fraction $b\sqrt{V}/a+b\sqrt{V}$ of the cooling is caused by the air velocity V, while all of the heating is caused by the current I. For the 3-μ tungsten wire the ratio between $b\sqrt{V}$ and a runs from 0 to about 2 at an air velocity of about 16 m/sec.

The factor 2 in the denominator of (16) stems from the square root in King's equation, while the same factor in the numerator of (17) stems from the second power of the current in Joule's law.

In the practical case of a tungsten wire, the temperature coefficient of the resistivity is approximately constant in the temperature range used. Thus

$$g_4 = \frac{r}{t} \approx \frac{R - R_a}{T - T_a} \cdot \tag{19}$$

Substituting (19) into (18) gives

$$\lambda = 1 - \frac{R_a}{R} \cdot \tag{20}$$

Substituting this equation into (14) for the time constant M results in

$$M = \frac{C(T - T_a)}{R_a I^2} \tag{21}$$

or

$$M = \frac{C(T - T_a)}{P} \cdot \frac{R}{R_a} \cdot \tag{22}$$

From (2) it is seen that the previously mentioned stability condition for cco, $g_4 g_5 < g_3$, is satisfied if M is positive. From (21) it follows that this condition is indeed satisfied.

Eq. (22) gives the time constant M as the ratio of the heat $C(T-T_a)$ accumulated in the wire and the static heat flow P with the correction factor R/R_a.

In constant-temperature operation, an increase in air velocity causes the factor P in (22) or the factor I^2 in (21) to increase because the parameters T and R are kept constant. Thus the time constant M of the hot wire alone is in this case inversely proportional to the electric power supplied to the wire or inversely proportional to I^2.

Using (5) and (6), (22) can be written

$$M = \frac{C}{a + b\sqrt{V}} \cdot \frac{R}{R_a} \cdot \tag{23}$$

This equation shows that at constant air velocity V the time constant is proportional to the hot-wire resistance R which depends on the operating temperature chosen.

III. Frequency Response of a Constant-Temperature-Operation Hot-Wire Anemometer

In a constant-temperature hot-wire anemometer the wire temperature is controlled by automatic adjustment of the current through the wire. To this end variations in the wire resistance which indicate the variations of the wire temperature are measured in a Wheatstone bridge circuit. The bridge output voltage is amplified, and the amplifier output current is supplied to the Wheatstone bridge, as shown in Fig. 2. The heating due to the current variations should balance the cooling due to the velocity variations.

In this section the response of the amplifier's output current to variations in air velocity will be discussed.

In the practical bridge arrangement, maximum efficiency has been obtained by making $R_2 \gg R_3 \gg R$, so that the full output current passes through the hot wire, and the voltage variations e across the hot wire are not carried over to R_1. Under these conditions, the branch voltages can be written

$$E_1 = \frac{R_1}{R_2} R_3 I, \text{ hence } e_1 = \frac{R_1}{R_2} R_3 i \tag{24}$$

and

$$E = RI, \text{ hence } e = rI + Ri. \tag{25}$$

If the bridge is balanced for static conditions,

$$R = \frac{R_1}{R_2} R_3 \tag{26}$$

and (24) can be written

$$e_1 = Ri. \tag{27}$$

In that case the bridge output voltage is found from (25) and (27) to be

$$e_b = e_1 - e = rI. \tag{28}$$

From this expression it is obvious that the bridge output is related only to variations in the hot-wire resistance.

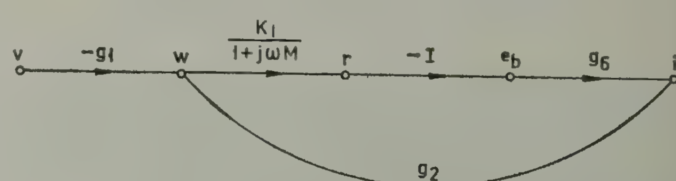


Fig. 5—Signal flow graph of cto anemometer in the idealized case of static bridge balance and zero transconductance of the amplifier for in-phase input signals.

Fig. 5 gives a signal flow graph of the cto anemometer for the above-mentioned idealized conditions. This figure differs from Fig. 4 in that the transfers $-I$ and g_6 are added, $-I$ being the transfer function of the bridge and g_6 being the transconductance of the amplifier. The latter is a function of the frequency. The amplifier's bandwidth is necessarily much larger than the frequency of the turbulence to be studied. The trans-

conductance g_6 can therefore, for a first approximation, be considered constant. The static loop gain of the feedback loop is

$$S = K_1 I g_6 g_2 = 2R\frac{\lambda}{1-\lambda}g_6 = 2R\left(\frac{R}{R_a}-1\right)g_6. \quad (29)$$

In the constructed instrument $S \gg 1$ so that the transfer function from v to i as derived from Fig. 5 can be written

$$\frac{i}{v} = \frac{g_1}{g_2}\frac{S}{1+S}\frac{1}{1+j\omega M'} \simeq \frac{g_1}{g_2}\frac{1}{1+j\omega M'} \quad (30)$$

with

$$M' = \frac{M}{1+S} \simeq \frac{M}{S}. \quad (31)$$

From (31) it is obvious that the time constant M' of the hot wire incorporated into the feedback loop can be made very small by providing a large loop gain S. Thus the cutoff frequency can be shifted from a few hundred cycles per second for the hot wire in an open circuit to several tens of kilocycles per second in the closed circuit.

From (30) it is seen that the static transfer from v to i is approximated by g_1/g_2 and thus is independent of the amplifier's transconductance g_6. It follows from Fig. 4 that the quotient g_1/g_2 is also equal to the quotient of r/v and r/i as given in (16) and (17), respectively, so that (30) can be written

$$\frac{i}{v} = \frac{I}{4V}\cdot\frac{b\sqrt{V}}{a+b\sqrt{V}}\cdot\frac{1}{1+j\omega M'}. \quad (32)$$

A few practical figures can now be derived. For a hot wire of 3 μ, with a length of 1 mm, the "cold" resistance R_a is about 10 ohms, and the cutoff frequency $1/(2\pi M)$ is about 200 cps at zero mean air velocity to 2000 cps at an air velocity of 40 meters per second (see Section VII).

To raise the cutoff frequency to 20 kilocycles per second, a static loop gain S of about 100 is necessary. A practical value of R_a/R is $\frac{2}{3}$, which makes $R = 15$ ohms.

Now the amplifier's necessary transconductance follows from (29):

$$g_6 = \frac{S}{2R\left(\frac{R}{R_a}-1\right)} \approx 7 \text{ mho}. \quad (33)$$

In the constructed instrument, values between 5 and 15 mhos are obtained, depending on the range of the output current chosen.

In the foregoing it has been presumed that the bridge is balanced in the static condition. In practice, however, there will always be a certain unbalance, *e.g.*, due to drift of the controlling amplifier or to a change in the mean air velocity.

Another complication of practical importance is caused by the nonzero transconductance of the amplifier for an in-phase input voltage, *i.e.*, a voltage between both input terminals and earth.

The influence of these undesired properties on the over-all transfer function will now be analyzed.

The output signal of the bridge follows from (24) and (25):

$$e_b' = e_1 - e\left(\frac{R_1}{R_2}R_3 - R\right)i - rI = \Lambda Ri - rI \quad (34)$$

with

$$\Lambda = \frac{R_1}{R_2}\frac{R_3}{R} - 1. \quad (35)$$

Λ is a measure of the static bridge unbalance. Comparing (28) and (34) shows that in the latter a term proportional to the bridge unbalance Λ is added.

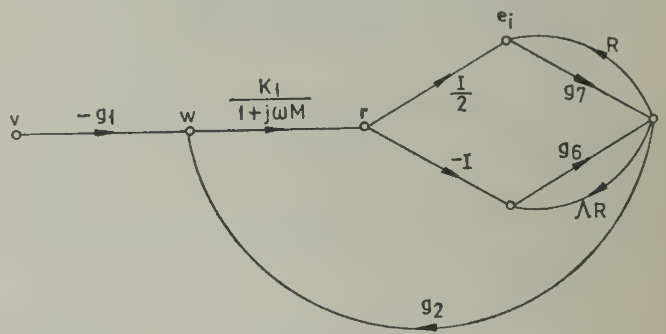


Fig. 6—Signal flow graph of cto anemometer in the general case of an amplifier with a finite rejection factor $f = g_6/g_7$ and a static bridge unbalance Λ.

In the signal flow graph (Fig. 6) this term is represented by a feedback path across the amplifier's transconductance g_6.

To find the influence of a transconductance for in-phase input signals of the controlling amplifier, make the in-phase signal component

$$\frac{e_1 + e}{2} = e_i \quad (36)$$

or, using (24), (25) and (35),

$$e_i = iR\left(1 + \frac{\Lambda}{2}\right) + \frac{rI}{2}$$

which at small values of Λ can be approximated by

$$e_i \approx iR + \frac{rI}{2}. \quad (37)$$

This equation is represented in the flow graph, Fig. 6, by the transfers R and $I/2$ from i and r to e_i. The transconductance of the amplifier for in-phase input signals is in Fig. 6 represented by g_7. Make

$$g_7 = \frac{g_6}{f} \quad (38)$$

where f is the rejection factor of the amplifier and the ratio of the gains for the anti-phase input signal e_b and the in-phase input signal e_i [10].

The static loop gain of the main feedback loop i-w-r-i in Fig. 6 is now

$$S' = \frac{S\left(1 - \frac{1}{2f}\right)}{1 - g_6 R\left(\Lambda + \frac{1}{f}\right)} \simeq \frac{S}{1 - q} \tag{39}$$

if

$$q = g_6 R\left(\Lambda + \frac{1}{f}\right). \tag{40}$$

From (39) it is seen that the loop gain S is increased or decreased by the additional loops in Fig. 6, depending on the sign of q.

The over-all transfer function from v to i is now

$$\frac{i}{v} = \frac{g_1}{g_2} \frac{S'}{1 + S'} \frac{1}{1 + j\omega M''} \simeq \frac{g_1}{g_2} \frac{1}{1 + j\omega M''} \tag{41}$$

where

$$M'' = \frac{M}{1 + S'} \simeq \frac{M}{S'} \tag{42}$$

or

$$M'' = M'(1 - q). \tag{43}$$

From (43) and (40) it is seen that a static bridge unbalance Λ and/or a finite rejection factor f of the amplifier for in-phase input signals have a considerable influence on the reduction of the hot wire's time constant. As g_6 and g_7 depend in a different way on the frequency, the rejection factor f in (40) is also frequency dependent.

IV. Stability Requirements

The idealized case of Fig. 5, where the amplifier is drift-free and not sensitive for in-phase input signals, has already been analyzed extensively by Eli Ossofsky [6].

The dynamic loop gain as derived from Fig. 5 is

$$S_d = -g_2 g_6 I \frac{K_1}{1 + j\omega M} \tag{44}$$

in which g_2 and K_1 are scalar factors, given by (11) and (4). In the case of a 3-stage amplifier with equal time constants τ per stage, the transconductance g_6 can be written

$$g_6 = \frac{g_0}{(1 + j\omega\tau)^3} \tag{45}$$

where g_0 represents the amplifier's transconductance at low frequencies. The influence of the drift correction amplifier (see Fig. 7) at zero and very low frequencies is discussed later on. Making

$$S_0 = g_2 g_0 I K_1 \tag{46}$$

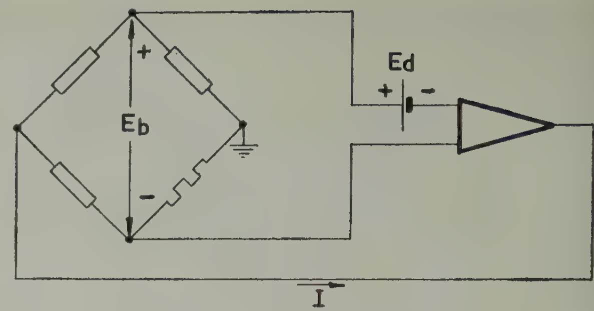


Fig. 7—If the amplifier's drift is equivalent to an input voltage variation E_d, a bridge unbalance voltage $E_b \sim E_d$ is generated by the feedback loop.

leads to

$$S_d = \frac{-S_0}{(1 + j\omega\tau)^3(1 + j\omega M)} \cdot \tag{47}$$

Eq. (47) is equivalent to (11b) of Eli Ossofsky [6]. To allow a static loop gain S_0 of 100, the ratio M/τ should be 500 or greater to prevent self-oscillation.

Turning to the case of Fig. 6, the loop gain can be derived to be

$$S_d' = \frac{1}{1 - q} \cdot \frac{-S_0\left(1 - \frac{1}{2f}\right)}{(1 + j\omega\tau)^3(1 + j\omega M)}$$

$$\simeq \frac{-S_0}{(1 - q)(1 + j\omega\tau)^3(1 + j\omega M)} \tag{48}$$

where q is given by (40).

Comparing (47) and (48) shows that the loop gain is increased or decreased by the additional feedback loops of Fig. 5, depending on the signs and magnitudes of f and q.

An increase in loop gain increases the bandwidth requirements for maintaining dynamic stability. To find the full dynamic requirements, it is necessary to study the frequency response (48), taking into consideration that q and f are functions of frequency.

Apart from dynamic instability, there is the possibility of "static" instability if the static loop gain $S_0/1-q$ becomes positive. Thus the condition for "static" stability is

$$q < 1 \tag{49}$$

which is safely met if both parts of q are small enough, *e.g.*,

$$g_6 R \Lambda < 0.3 \tag{50}$$

and

$$\frac{g_6 R}{f} < 0.3. \tag{51}$$

Sources of bridge unbalance Λ are drift of the controlling amplifier and changes in the mean air velocity.

Drift of the amplifier is equivalent to a drift voltage source put in series with the input leads of the amplifier,

for instance, E_d in Fig. 7. Because of the feedback, this voltage is balanced by a bridge output voltage

$$E_b = E_d \frac{S}{1+S} \simeq E_d. \tag{52}$$

Because

$$E_b = \Lambda E = \Lambda IR \tag{53}$$

(50) can be written

$$\frac{g_6 E_d}{I} < 0.3. \tag{54}$$

Practical values $g_6 = 10$ mho at $I = 30$ ma and substituted in (54) give the amplifier's drift requirements

$$E_b = E_d < 1 \text{ mv}. \tag{55}$$

In the long run this requirement can only be met if use is made of a chopper stabilized amplifier [12], as shown in Fig. 8. The chopper dc amplifier in this arrangement lifts the static gain of the amplifier by about a factor 500, thus reducing the drift of the amplifier by about the same factor.

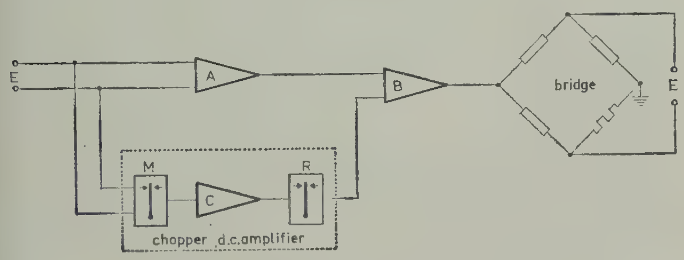


Fig. 8—A chopper dc amplifier is added to the circuit to meet the drift requirements. The blocks A and B represent wide-band dc amplifiers. The chopper dc amplifier consists of the modulator M, the ac amplifier C and the synchronous rectifier R.

Changes in the mean air velocity cause changes in the mean bridge current I, which maintain the hot wire at its correct resistance. Values found in practice are about 20 ma at zero air velocity to 100 ma at an air velocity of 40 m/sec. With an average transconductance in this range of $g_6 = 10$ mhos, the variation in the amplifier's input (or bridge output) voltage needed to bring about this output current variation is 6 mv, which is higher than the figure quoted in (55). In the arrangement shown in Fig. 8, however, the static gain is increased by about a factor 500, so that the actual input voltage variation needed is not 6 mv, but only 12 μv. Thus the chopper stabilizer offers the additional advantage of correcting automatically for changes in the mean air velocity.

The amplifier's rejection requirements are found from (52):

$$f > \frac{g_6 R}{0.3} \tag{56}$$

Putting $R = 15$ ohms and $g_6 = 10$ mhos into (56) gives

$$f > 500. \tag{57}$$

The rejection requirements for the chopper dc amplifier are much more exacting because at zero and very low frequencies it lifts the gain of the amplifier by about a factor 500. Thus the rejection factor of the chopper amplifier should exceed $500 \times 500 = 250{,}000$.

V. Signal-to-Noise Ratio

It seems to be commonly held [6] that the SNR of comparable hot-wire anemometers is less favorable in the case of those having constant temperature operation since these require, for reasons of stability, a considerably larger bandwidth in the controlling amplifier than do those employing constant current operation.

It is the authors' opinion that this statement in its above-given general form is not valid; it is certainly not only the bandwidths of the different controlling amplifiers which determine the SNR at the instrument output terminals.

All the noise (thermal, shot, partition, secondary emission) in a certain frequency band can effectively be replaced by a single noise source at the input of the amplifier. The only basic difference between the setup of a cto and a cco hot-wire anemometer is, then, that the former has a feedback loop and the latter has not. However, it is well known that feedback does not influence the SNR for noise sources at the input of a system. If, therefore, the quality of the first stage of the controlling amplifiers and the over-all bandwidth of the measuring setup are equal for the cto and the cco systems, their SNR will also be equal.

If in the case of comparable instruments (*i.e.*, instruments designed for measuring turbulence up to the same maximum frequency) the bandwidth of the controlling amplifier is larger for the cto hot-wire anemometer than for the cco instrument, it goes without saying that the amount of high-frequency noise in the case of cto is larger at the output of the controlling amplifier. However, at the output of the hot-wire anemometer as a whole the additional noise is cut off in the low-pass output filters that determine the over-all bandwidth.

VI. Method of Measuring the Over-all Transfer Function

It is extremely difficult, if not impossible, to measure the transfer function i/v directly by varying the air velocity. It is possible, however, to measure this transfer function in an indirect way by introducing an electrical disturbance into the control circuit [6].

The measuring setup is shown in Fig. 9. An external current i_e from a high-impedance source is introduced into the circuit, and the response of the amplifier output current i_a is measured. For the idealized case that $q = 0$ (no static bridge unbalance, infinite rejection of in-phase inputs), the transfer function is found from the signal flow graph of Fig. 10,

$$\frac{i_a}{i_e} = -\frac{S}{1+S} \cdot \frac{1}{1+j\omega M'} \simeq -\frac{1}{1+j\omega M'} \tag{58}$$

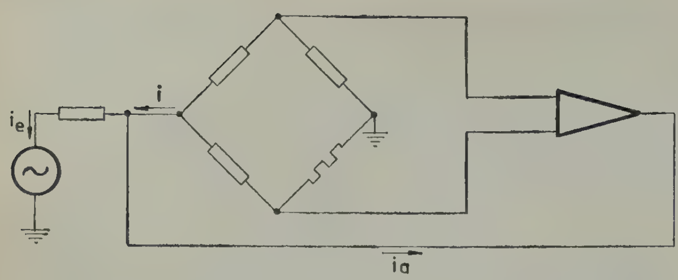


Fig. 9—A sinusoidal current i_e is injected into the circuit to find the frequency response of the cto anemometer.

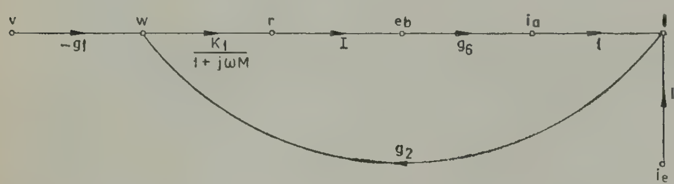


Fig. 10—Signal flow graph of the circuit given in Fig. 9 for $q=0$; that is at zero bridge unbalance and infinite rejection of in-phase input signals.

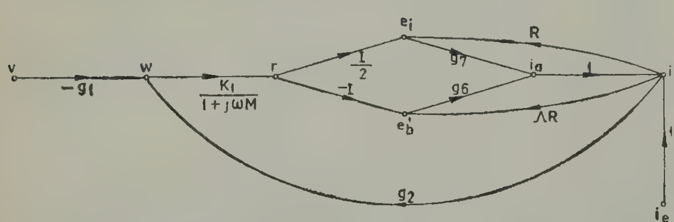


Fig. 11—Signal flow graph of the circuit of Fig. 9 in the practical case of a static bridge unbalance Λ and an amplifier with a finite rejection factor $f=g_6/g_7$.

where S and M' are given by (29) and (31), respectively.

Comparing (30) and (58) one notices that the transfer functions i/v and i_a/i_e differ only by a scalar factor. Thus to find the transfer characteristics of i/v one could instead measure i_a/i_e which is more attractive since i_e can easily be varied sinusoidally or stepwise.

In the case of a finite rejection factor f and a static bridge unbalance, the signal flow graph of Fig. 11 is valid with the transfer function

$$\frac{i_a}{i_e} \simeq \frac{1}{1-q}\left(q - \frac{1}{1+j\omega M''}\right). \tag{59}$$

The response consists then of two parts: one part is a proportional response, the other part is governed by the transfer function $1/(1+j\omega M'')$. The proportional part can be either positive or negative depending on the sign of q.

Both the sign and the magnitude of the proportional part q can be found simply by varying i_e stepwise. The step response of the proportional part is another step; the step response of the other part is a first-order exponential function with a time constant M''. The ratio of the static values of the proportional and the exponential parts of the total step response is then equal to $-q$.

Thus, in the case where $q \neq 0$, this method can likewise be used to find the transfer characteristic of i/v.

VII. Results of Measurements and Calculations

Eq. (21) can be rewritten as follows:

$$M = \left(\frac{\pi}{4}\right)^2 \cdot d^4 \cdot \beta \cdot \gamma \frac{T - T_a}{\rho_a \cdot I^2} \cdot \tag{60}$$

Here

d = the diameter of the hot wire.
β = the density of the hot-wire material.
γ = the specific heat of the hot-wire material.
T = the hot wire temperature.
T_a = the air temperature.
ρ_a = the specific resistance of the hot-wire material at the temperature of the air.
I = the heating current.

From (60) it is possible to calculate the time constant. For a tungsten wire with diameter 3 μ, operated under the following conditions: $I=20$ ma, $T-T_a \approx 100$°C, $V=0$ m/sec, M is found to be about 0.6 msec.

Another way of determining M is by direct measurement. This can be done in the same bridge circuit as given in Fig. 2. In this case, however, the loop is not closed. In the setup of Fig. 12 the amplifier A_1 is only used as a means to amplify the voltage e_b, which in (28) was found to be proportional to the variations r in the hot-wire resistance. Another amplifier A_2 is used to amplify the voltage e_1, which is a measure of the current i [see (24)].

At static bridge balance and complete rejection of in-phase signals, the signal flow graph of Fig. 13 is valid, from which is found

$$\frac{e_4}{i} = -\frac{SR_4}{1+j\omega M} \cdot \tag{61}$$

At static bridge unbalance and incomplete rejection of in-phase input signals, the signal flow graph of Fig. 14 is valid, which results in an over-all transfer:

$$\frac{e_4}{i} = -R_4\left(\frac{S}{1+j\omega M} - q\right).$$

In the case of an open circuit, however, it is easy to compensate the incomplete rejection of in-phase signals by introducing a static bridge unbalance to make $q=0$.

The output voltages e_4 and e_5 in the setup of Fig. 12 were compared on an oscilloscope. Thus the ratio r/i was measured as a function of the frequency. This resulted in the locus of Fig. 15, which indeed shows the behavior of a first-order system as was found in Section II.

The time constant M as derived from Fig. 15 is about 0.8×10^{-3} seconds. This is in fair agreement with the value calculated above from (60), especially if one takes into consideration that a small amount of dust on the wire has a considerable influence on the time constant. Moreover, owing to the fourth power of the wire diameter d occurring in (60), an error of 10 per cent in d brings about a change of about a factor 1.5 in the time constant M.

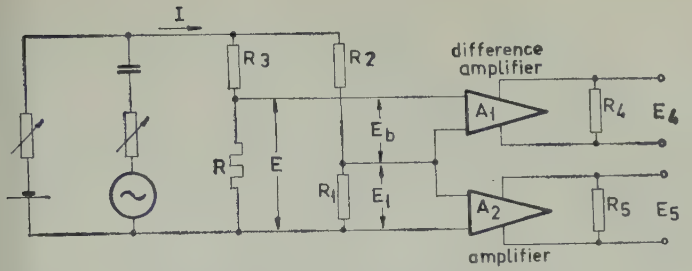


Fig. 12—Arrangement for measuring the hot wire's frequency response.

Fig. 13—Signal flow graph of the circuit of Fig. 12 under idealized conditions.

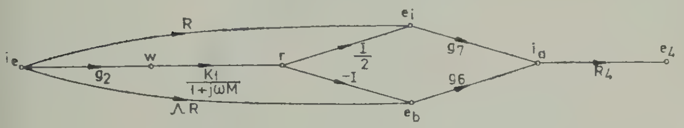


Fig. 14—Signal flow graph of the circuit of Fig. 12 under practical conditions.

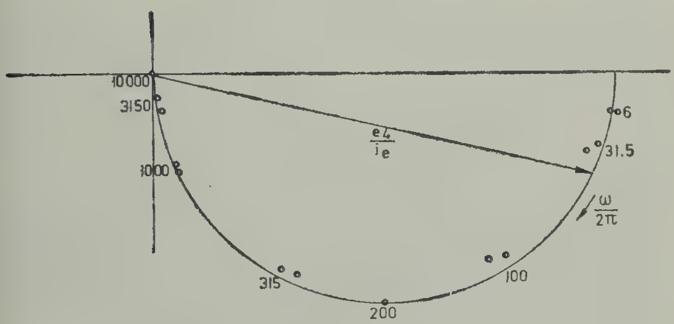


Fig. 15—Measured locus of the hot wire's frequency response.

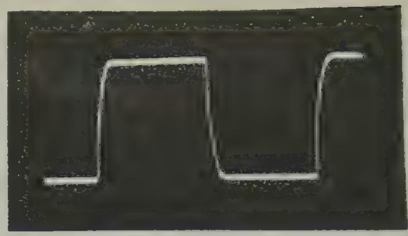
The scatter in the results of the measurements could in part be due to the fact that several burnouts of the hot wire prevented the various measurements from being carried out with one wire.

An effective way of testing the over-all performance of a cto hot-wire anemometer is by introducing a step signal i_e into the setup of Fig. 9.

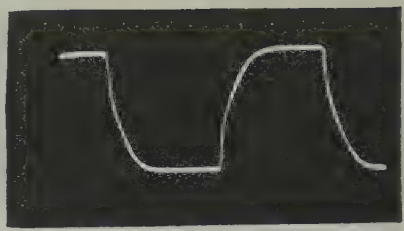
Fig. 16(a) and 16(b) give the response of i_a to a square wave disturbance in i_e with frequencies of 3000 and 10,000 cps, respectively; both figures are for zero air velocity. It is seen that the response of i_a consists of two parts, the amplitude ratio of these two parts being equal to q [see (59)]. From Fig. 16(b) one finds $q=-0.8$.

The drift correction amplifier maintains the static unbalance of the bridge well within ± 100 μv, so that the contribution of Λ to q [see (40)] is negligible. The value of f is now found from (40) by putting $g_6=10$ mhos and $R=15$ ohms at $q=-0.8$ and $\Lambda=0$:

$$g = \frac{g_6 R}{q} = -190.$$



(a)



(b)

Fig. 16—(a) Response of i_a to a square-wave disturbance in i_e with a basic frequency of 3000 cps. Zero air velocity. (b) Same as (a) but with a basic frequency of 10,000 cps.

Thus the absolute value of f does not meet requirement (57), but since the sign is negative, f causes negative feedback. In a future design an attempt should be made to improve this figure.

Making $R=15$ ohms, $R/R_a=1.5$, $g_6=10$ mhos and $q=-0.8$ in (29) and (39) one finds

$$S = 150 \quad \text{and} \quad S' = 80.$$

By means of (42) it is now possible to calculate M''. From the measured value of $M=0.8$ msec one finds $M''=10$ μsec, which is in good agreement with the time constant of the exponential part of Fig. 16(b) [see (59)].

From $M''=10$ μsec we can calculate the 3-db point to be 16×10^3 cps. Direct measurement of the 3-db point by means of sinusoidal variations in i_e yielded a value which was about 20×10^3 cps.

In the tests in which Fig. 16(a) and 16(b) give the response of the amplifier current i_a, the air velocity chosen was zero since there is then only one kind of signal entering the control circuit.

To find the over-all performance of a hot-wire anemometer measurements under various operating conditions are necessary because the time constant of the wire depends on the air velocity. At air velocities not equal to zero, i_e should be made sufficiently high to be well above the signal level of the turbulence.

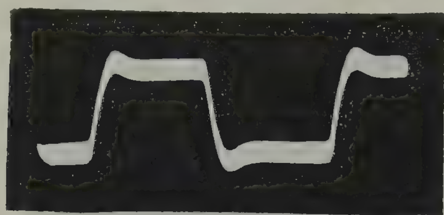
It was found that the constructed cto hot-wire anemometer showed a drop of 1 per cent in the modulus of the transfer function i/v at a frequency of about 20×10^3 cps under the following conditions:

$$V \approx 3 \text{ m/sec}$$

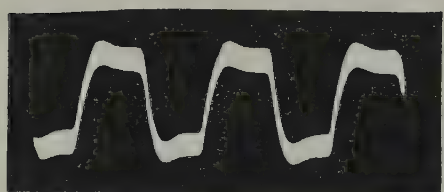
$$\frac{R}{R_a} = 1.57$$

$$I = 65 \text{ ma}.$$

Fig. 17(a) and 17(b) gives a picture of the response of i_a to a square-wave disturbance in i for $V \neq 0$. These figures are partly blurred due to the turbulence signal. It should be noted that Fig. 17(a) and 17(b) shows a certain overshoot, which is not present in Fig. 16(a) and 16(b). This phenomenon is to be ascribed to a lower value of M [see (21)], and the influence of M on the dynamic loop gain according to (48).



(a)



(b)

Fig. 17—(a) Response i_a to a square-wave input i_e with a basic frequency of 10,000 cps. Air velocity not zero. (b) Same as (a) but with a basic frequency of 20,000 cps.

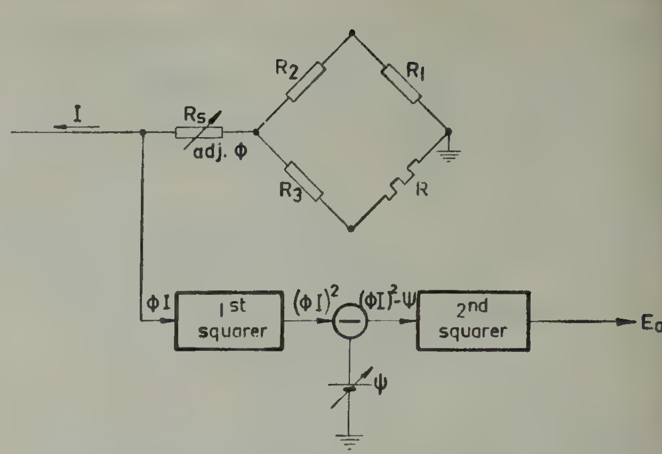


Fig. 18—Block diagram of linearizer. The output E_0 is proportional to the measured air velocity.

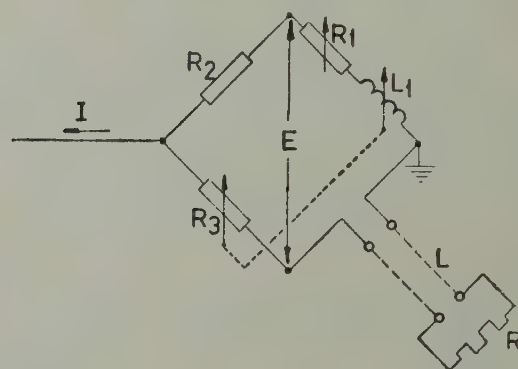


Fig. 19—Bridge arrangement.

VIII. Linearizing

From King's equation (5) and the power equation (6) it follows that the relation between the bridge current I and the air velocity V will be

$$V = \left\{ \frac{\dfrac{R}{T - T_a} I^2 - a}{b} \right\}^2 \tag{62}$$

or

$$V = \{(\Phi I)^2 - \Psi\}^2 \tag{63}$$

with

$$\Phi = \sqrt{\frac{R}{b(T - T_a)}} \tag{64}$$

and

$$\Psi = \frac{a}{b}. \tag{65}$$

The parameters Φ and Ψ depend on the hot-wire dimensions and the hot-wire temperature T.

The block diagram of the linearizer is given in Fig. 18. A variable resistor R_s in series with the bridge is used to adjust the parameter Φ. Two squarers perform the squaring operations of (63). The parameter Ψ is introduced in a substracting unit between the two squarers.

During a test the parameters Φ and Ψ were adjusted so as to obtain maximum output at an air velocity of 40 m/sec and quarter-scale output at 10 m/sec. Succeeding measurements showed the output voltage E_0 to be proportional to the air velocity within 2 per cent.

IX. Construction

The starting point for the design and construction was to make an instrument for the kilocycle range, provided with linearizing and suitable for use by people without experience in electronics.

The restricted frequency range allowed the hot wire to be connected to the bridge by means of a connecting cable. The other elements of the bridge (see Fig. 19) are located in the instrument. The equilibrium value of the hot wire's operating resistance is read directly from the positions of two switches in the branches R_1 and R_3 respectively. R_3 is for coarse adjustment in 3 steps and R_1 for fine adjustment in 11 steps. The total range of R is from 4.5 to 44 ohms.

The series self-inductance L of the hot-wire probe leads and connecting cable is corrected by means of a self-inductance L_1 in series with R_1. As

$$\frac{L_1}{L} = \frac{R_1}{R} = \frac{R_2}{R_3} \tag{66}$$

and L and R_2 are constant; the value of L_1 is related only to that of R_3.

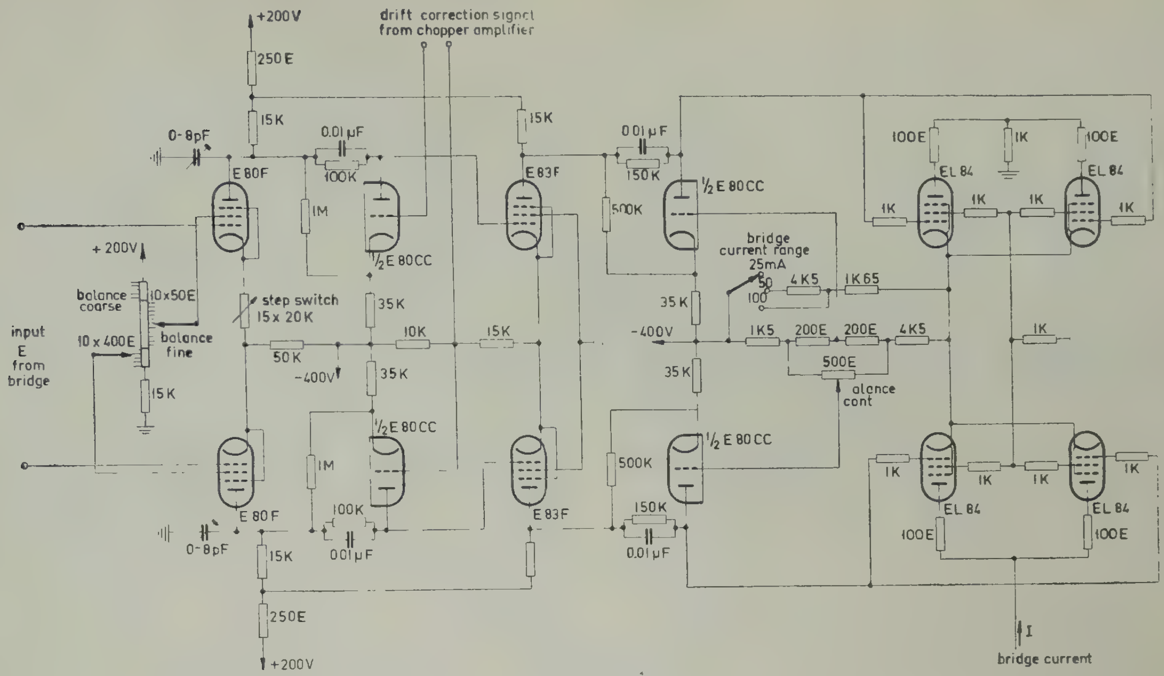


Fig. 20—Control amplifier.

The dc amplifier (Fig. 20) is a further development of Ossofsky's amplifier [6]. To obtain a large rejection factor for in-phase signals, the input tubes are equalized by means of an adjustable resistance in series with one of the cathodes. In a more advanced design, the rejection factor could be considerably improved without any equalizing means at all by applying the technique described by Klein [10]. Initial balance is adjusted by means of two tap switches in the screen grids supply of the first stage and by a potentiometer in the voltage divider between the second and the output stages. The bridge current range is chosen by means of an adjustable resistor in the common cathode circuit of the output stage. The drift correction signal from the chopper amplifier is injected in one of the Mezger type voltage dividers [11] between the first and the second amplifier stages.

The chopper amplifier is shown in Fig. 21. The input filter has two sections. The first section is symmetrical and serves to attenuate in-phase ac components, the second to attenuate anti-phase ac components. To obtain a good rejection factor both statically and dynamically for in-phase signals, the amplifier has a differential input stage. To prevent insulation currents setting up a static difference input signal across R_1 and R_2, a system of guarding (not shown in Fig. 21) was needed in the chopper circuit.

The output chopper rectifier is followed by a filter for reducing the ripple voltage.

The entire chopper amplifier circuit now contains three low-pass filter sections, *i.e.*, two at the input and one at the output. To prevent instability in the closed loop due to the phase lag of these filters, the time constant of the output filter was increased to 800 seconds.

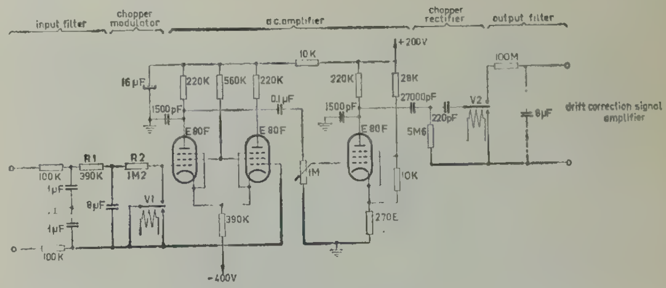


Fig. 21—Drift correction amplifier.

An electronic galvanometer is permanently connected to the bridge output. Its circuit is a simplified version of that of the drift correction amplifier. It serves as a continuous indicator of the bridge unbalance voltage but does not load the bridge, and the indicator is not damaged by overload, *e.g.*, in the case of hot-wire burnout.

The block diagram of the linearizer is shown in Fig. 18. The electronic circuit is given in Fig. 22. The squarers are of the diode function generator type [12].

Fig. 23 shows the complete anemometer without hot-wire probe. The upper panel contains the control amplifier, the chopper amplifiers for the galvanometer and the drift correction, the linearizer, an attenuator and a pre-amplifier for the ac output signal. The center panel contains low-pass filters, an integrator circuit for studying macro-scale eddies, a differentiator circuit for studying micro-scale eddies [13], [14], and a rms voltmeter.

The lower panel contains supply units for the stabilized dc supply, not only of anode and screen voltages, but also of the filament currents of several critical tubes in the circuits.

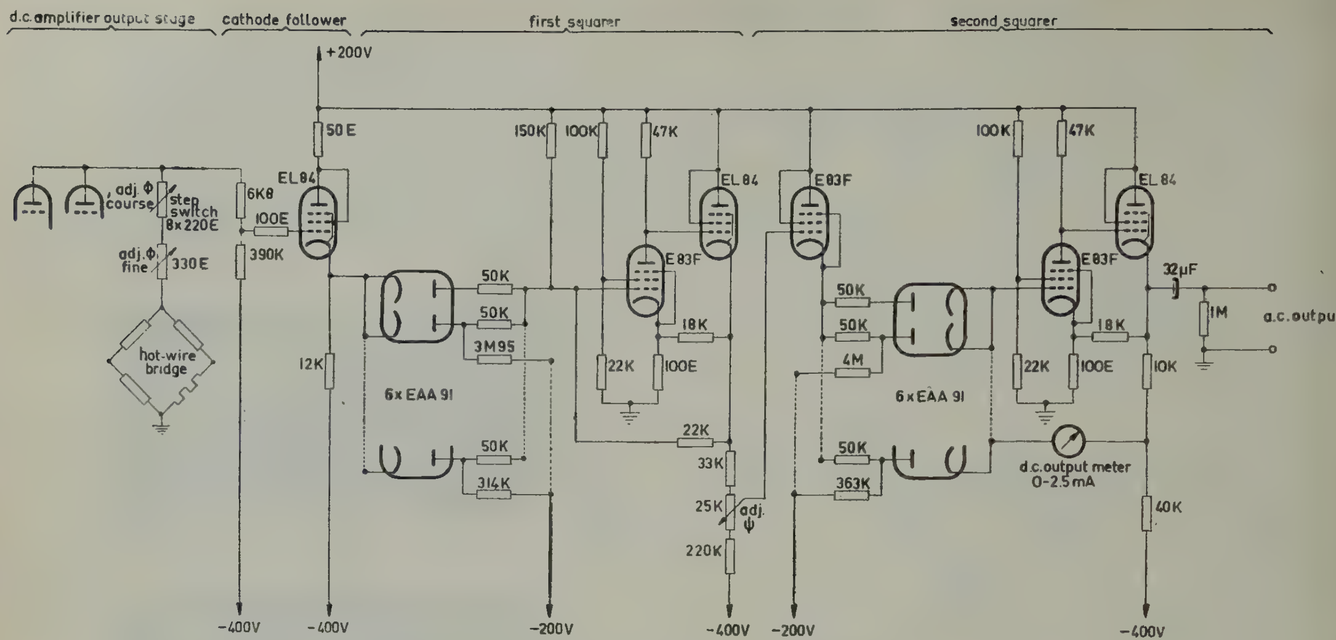


Fig. 22—Linearizer.

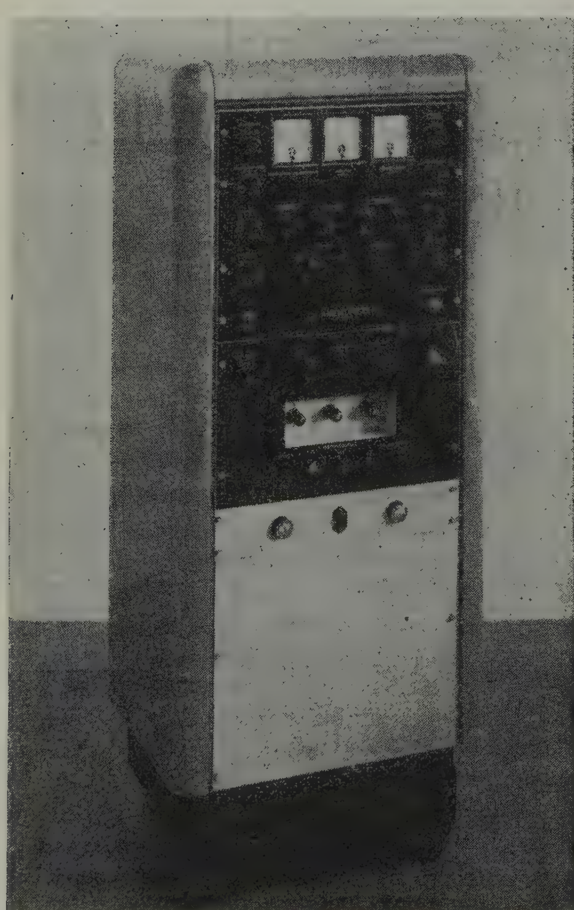


Fig. 23—The constructed constant-temperature operation hot-wire anemometer.

X. Other Types of CTO Hot-Wire Anemometers

Various ways of controling the resistance of the hot wire other than that described above are possible. The main features by which these systems can be distinguished are: 1) the way in which the hot-wire resistance is measured, and 2) the method by which the hot wire is heated.

In the system which is the subject of this paper, the measuring of the resistance and the heating of the hot wire are both accomplished by means of a dc current. This has a drawback in that the heating current interferes with the measuring signal when the bridge is not well balanced.

However, in a system which measures the hot-wire resistance by means of a RF current and which heats the hot wire with a dc current (or reversed: measuring with dc and heating with RF current) as has been suggested by Ziegler [4] one can obtain a system in which the heating does not influence the measurement. Figs. 24 and 25 give two possible systems; Figs. 26 and 27 give their respective signal flow graphs.

It is seen that the nonessential loops can in this way be eliminated. However, the number of elements in the control loop is increased, each element adding to the total phase lag in the loop.

It remains to be seen whether the advantages of a system where measurement and heating of the hot-wire resistance do not influence each other are a sufficient compensation for the stability problems that will occur as a result of the additional phase lags in the control loop.

Other possible systems would be those using radio frequency currents both for measuring and for heating.

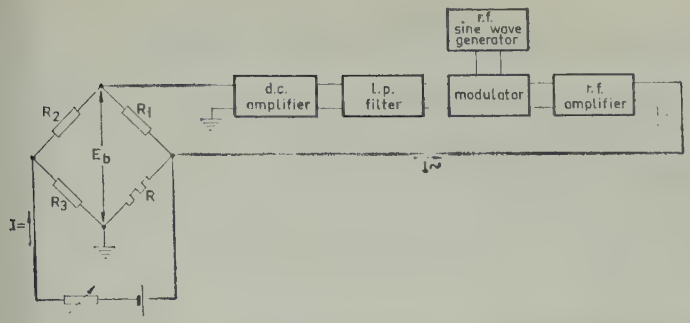


Fig. 24—CTO anemometer with dc bridge supply and ac heating current supply.

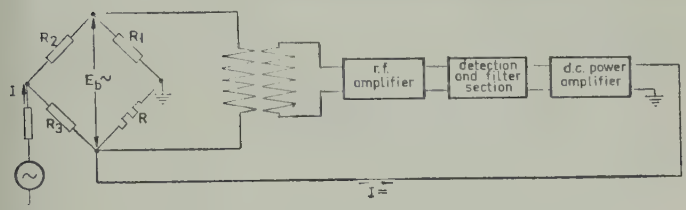


Fig. 25—CTO anemometer with ac bridge supply and dc heating current supply.

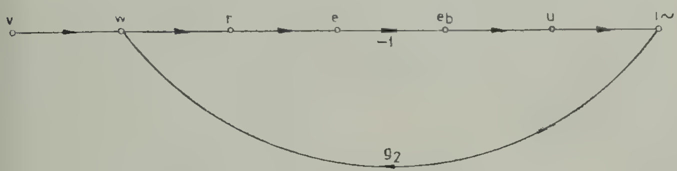


Fig. 26—Signal flow graph of the circuit of Fig. 24.

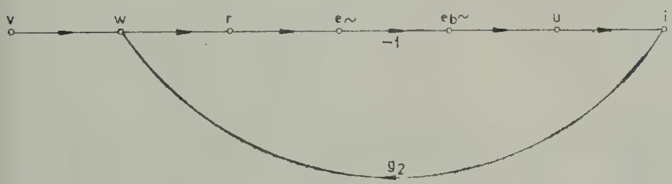


Fig. 27—Signal flow graph of the circuit of Fig. 25.

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Correction

Herbert Friedman, author of "Rocket Observations of the Ionosphere," which appeared on pages 272–280 of the February, 1959 issue of Proceedings has requested that the following correction be made to his paper.

Eq. (2) on page 275, should read:

$$N_2+O^+\rightarrow NO^+ + N.$$

IRE Standards on Antennas and Waveguides: Waveguide and Waveguide Component Measurements, 1959*

59 IRE 2. S1

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Section 1—General

1.1 Introduction

THIS STANDARD is concerned with measurements of the quantities which characterize a waveguide or waveguide component and the associated electromagnetic fields. The term "waveguide," as used here, is a generic term which includes transmission lines and uniconductor waveguides as special cases.

The measurements are described in general terms. For the specific details of procedure and equipment, the reader is directed to numerous references (Sec. 6), which form an indispensible part of this Standard.

Definitions of terms employed in this Standard are given in Sec. 5. These definitions are presented both to insure precision of meaning with regard to the measured quantities and to clarify possible conceptual differences in the usage of terms common to both microwave and low-frequency circuit practice. As an example of the latter, the term "match" has different significance in the two usages.

Section 2 is concerned with measurements in a waveguide having uniform characteristics. Sections 3 and 4 are concerned with components which are connected to or located in waveguides.

This Standard is limited to measurements in linear, reciprocal systems, and unless otherwise stated to single frequency sources. For example, it does not apply to special devices such as nonreciprocal ferrite components and crystal detectors.

1.2 Measurement Techniques

In general, the equipment required for waveguide measurements includes a signal source of the appropriate frequency and a suitable detector. In addition depending on the nature of the measurement being made, equipment such as attenuators, slotted lines, impedance bridges, sampling devices, and terminations may also be required. The exact form of this equipment will vary depending upon the nature of the waveguide, the frequency range, and the particular method used for the measurement.

Regardless of the equipment used, there are certain basic precautions which must be taken if the measurements are to be meaningful. The signal supplied by the source should be stable and free of harmonics and other spurious components and should be free of undesired modulation. In many cases, this requirement may be relaxed if the detector is selective enough so as not to respond to the undesired signals. Leakage into a detector either directly from the signal source, through undesired paths in the component, or from external sources should be guarded against. If the detector is used at other than a constant level, it is often necessary to know accurately the response law of the detector. In connecting the measuring equipment to the device being measured, care should be taken to minimize unwanted discontinuities or reflections. If high precision is required, such discontinuities as do exist must be separately determined and accounted for in the data reduction. In certain kinds of measurements, it is desirable that the impedance of the source be matched to the characteristic impedance of the waveguide being used. For the case of balanced transmission line systems, care must be taken that the probe and detector respond only to the balanced component of the existing fields. This is a special case of the more general problem involving measurements in waveguides which can support multimode transmission, and it is obvious for such cases that the detector should respond only to the mode of interest.

Section 2—Measurements in Waveguides [1], [2]

2.1 Measurement of Voltage

(a) In transmission lines carrying a TEM mode, it is possible to measure the total voltage across the line (*i.e.*, the line integral, in the transverse plane, of the electric field strength) at frequencies below a few hundred mc with a reasonable degree of accuracy by using vacuum tube voltmeters or crystal diode voltmeters which are available commercially. In making this type of measurement, care should be taken to insure that the voltmeter does not unduly load the line. In addition, the voltmeter leads should not couple to any field other than that being measured.

(b) In a balanced transmission line system it is necessary that a balanced type of voltmeter be used. This may consist of an unbalanced voltmeter with a balance-to-unbalance transformer of known ratio. A second method of measuring a balanced voltage is to use a quarter-wave shunt stub of balanced transmission line with an ammeter shorting the line at the far end. The voltage across the line is equal to the current times the characteristic impedance of the stub line.

(c) In uniconductor waveguides, total voltage is not usually of interest. However, measurement of electric field strength is frequently of interest (see Sec. 2.3).

2.2 Measurement of Current [1], [3]

(a) In a transmission line carrying a TEM mode, the total current flowing in either conductor can be measured with a reasonable degree of accuracy at frequencies below several hundred mc with a thermo-couple type ammeter. At frequencies much lower than this, a thermo-ammeter is sometimes used.

(b) In a balanced transmission line system, the total current may be measured by inserting ammeters in series with each conductor or by using a single ammeter and a suitable coupling loop such that the loop couples only to the balanced component of the current [4]. If unbalanced currents are present, as may be evidenced by a difference in the meter readings on the two sides of the line or by other methods (see Sec. 2.12), it will be necessary to locate and remove the sources of the unbalance before proceeding with the measurement.

(c) In a uniconductor waveguide, the definition of total current has no practical importance. However, the local current density at a point on the surface can be determined by measuring the magnetic field strength at that point (see Sec. 2.4).

2.3 Measurement of Electric Field Strength [5]

The component of the electric field in a specified direction at a point in space is usually obtained by measuring the voltage induced in a small linear probe oriented in the specified direction. Care must be taken that the transmission line leading to the probe does not itself affect the field nor the probe pickup. It is important that the amount of power absorbed or scattered by the probe be small compared to the total power in the guide.

Although it is possible to make absolute electric field strength measurements by calibrating the probe, relative measurements are usually of most interest.

2.4 Measurement of Magnetic Field Strength [5]

The component of the magnetic field in a specified direction at a point in space is usually obtained by measuring the voltage induced in a small loop type probe oriented in the specified direction. Care must be taken that the transmission line leading to the probe does not itself affect the field nor the probe pickup. It is important that the amount of power absorbed or scattered by the probe be small compared to the total power in the guide.

It should be noted that a loop type probe will always have a certain amount of response to the electric field as well as to the magnetic field. This response can be minimized by keeping the loop and the associated meter balanced and by keeping sizes as small as possible. Balance can be checked by physically reversing the loop.

When measuring the magnetic field strength on a metallic surface, a small rectangular slot in the surface is sometimes used as the probing device. For this case, the voltage across the slot is measured.

Although it is possible to make absolute magnetic field strength measurements by calibrating the probe, relative measurements are usually of most interest.

2.5 Power Measurements [6]–[9]

(a) In a matched lossless transmission line carrying a TEM mode, the power absorbed by the load (which is equal to the incident power) can be determined from the characteristic impedance (Z_0) of the line and either the total voltage (V) or the total current (I). If the line is mismatched, the power (P) absorbed by the load can be computed from the value of either of these quantities at both a maximum and minimum.

$$P = \frac{V_{\max} V_{\min}}{Z_0} = I_{\max} I_{\min} Z_0.$$

The absorbed power is sometimes measured by the three meter method in which either voltage or current readings are taken along a line at three appropriately spaced points [10].

(b) In a matched lossless waveguide, the incident power can be determined by measuring the power dissipated in the matched load or by extracting and measuring a known fraction of that power. If the waveguide is mismatched, the incident power can be determined by extracting, and measuring a known small fraction of the forward traveling wave by means of a directional coupler.

(c) There are several common methods of measuring the power in a matched load.

1. Measurement of resistance change. This method includes the use of devices such as bolometers and thermistors [7].
2. Calorimetric methods [7].
3. Thermo-electric (*e.g.*, thermo-couple) methods [7].

2.6 Measurement of Waveguide Wavelength [11]–[13]

A common means of measuring waveguide wavelength employs a movable probe for observing the standing wave along a length of the waveguide which is terminated by a mismatched load. The use of a short-circuit termination as the mismatched load will improve the accuracy of the measurement. A half wavelength in the waveguide is equal to the distance between adjacent minima of the standing wave. In the case of completely enclosed waveguides it is necessary to provide a nonradiating longitudinal slot for the movable probe. Commercial slotted sections are available for most standard waveguides.

An equivalent procedure uses a fixed probe and a movable short circuit to position successive minima of the standing wave at the probe.

An alternate method employs a cavity arrangement wherein the waveguide under measurement is terminated at both ends by short circuits, one of which is adjusted for two successive resonances of the cavity.

For waveguides of simple cross section it is possible to compute the waveguide wavelength from a measurement of the frequency and a knowledge of the dimensions of the waveguide cross-section.

2.7 Measurement of Attenuation Constant [14]

The attenuation constant is the real part of the propagation constant and is the attenuation per unit length, expressed in nepers/unit length.

(a) The methods of measuring attenuation constant fall into two main groups, one based on transmission and one on reflection procedures.

(b) Transmission type procedures involve the measurement of the insertion loss of a known length of the waveguide under test by the methods of Sec. 3.1.

(c) Reflection type measurements require a knowledge of the magnitude of the input reflection coefficient of a known length of waveguide terminated in a load of known reflection coefficient [15].

(d) A variation on the reflection type measurement, particularly useful for waveguides of unusual cross-

section, employs the measurement of the Q of a cavity consisting of the length of waveguide under test, short-circuited at both ends (see Sec. 3.8) [16]. The end losses must be small or separable.

2.8 Measurement of Input Impedance [17], [18]

It is necessary to distinguish between input impedance and normalized input impedance. Input impedance is generally significant in transmission lines carrying a TEM mode; in uniconductor waveguides it is practicable to measure normalized input impedance. Input admittance is the reciprocal of input impedance.

At frequencies below several hundred mc, lumped constant bridges are useful in determining impedance [17]. At higher frequencies, impedance may be found by measuring the normalized impedance in a transmission line of known characteristic impedance carrying a TEM mode.

2.9 Measurement of Normalized Input Impedance [18]–[21]

The quantities Z', the normalized input impedance, Y', the normalized input admittance, and ρ, the complex voltage reflection coefficient, are simply related in the following manner

$$Z' = \frac{1}{Y'} = \frac{1+\rho}{1-\rho}.$$

Another associated quantity, the standing wave ratio (S) is related to the magnitude of ρ as

$$S = \frac{1+|\rho|}{1-|\rho|}.$$

Methods of measuring normalized input impedance may be classified under distributed constant bridge methods, standing wave ratio techniques, and procedures involving the separation of incident and reflected traveling waves.

(a) One class of distributed constant bridges consists of comparison devices which either directly compare the unknown with a known normalized impedance, or which effect the comparison through variable ratios [18], [21]–[24]. Another type of distributed constant bridge determines the complex ratio of the electric and magnetic field strengths [25], [26]. A number of commercially available bridges fall into these categories.

(b) Standing wave ratio techniques include a number of methods based upon the sampling of the field variation in a waveguide terminated by the unknown impedance. The most common of these procedures employs a sliding probe [20]. Other methods include the use of multiple fixed probes [27], or sweep methods such as frequency sweeping or the Chipman line procedure [28], [29].

(c) Procedures involving the separation of incident and reflected traveling waves, such as those employing directional couplers, yield directly the reflection coefficient magnitude only [30]. Phase information is sometimes obtained by the introduction of a known (usually capacitive) susceptance in a known location [31], [32].

2.10 Measurement of Dielectric Voltage Breakdown [33], [34]

(a) In a waveguide, it is possible to measure the dielectric breakdown of a particular region if its power-handling ability is less than that of the remainder of the line under test. Conditions affecting the lowest power level at which corona or arcing (breakdown) first appears include pressure, temperature, frequency, ionization, extraneous material, and magnitude of standing wave at the location of the breakdown.

(b) To measure power capacity, breakdown is induced by raising the applied power or lowering the ambient pressure, whichever is more convenient, and noting the pressure, temperature, and load conditions at breakdown. Breakdown may be observed by various methods including sight, sound, increased reflection, change in generator output, changes in transmission properties, or heating effects. Power capacity under most other conditions of pressure, temperature, and load reflection may then be computed. If an accurate measurement of the applied peak power is not available, due to irregular waveform, etc., a comparison may be made between a component of known breakdown, and the part under test [35]. Repeatability of results may be improved by irradiation.

(c) The variation of high-frequency breakdown with pressure and distance may be computed by a simple rule [33], [36] except under conditions of relatively low pressure and small distances, where the observed power capacity will be significantly greater than the computed value [33].

(d) Under short-pulse high-power conditions, it is possible to obtain stable states of "breakdown probability" less than unity, in which only a certain percentage of applied pulses break down. The peak power for zero probability of breakdown is known as the "threshold" and is generally computed by extrapolation from observed percentages of pulses breaking down at higher powers. The entire range between the threshold and unity breakdown probability is known as the "over-voltage" region. For convenience in testing, a point within the over-voltage region corresponding to a fairly low probability (one or two arcs per minute) is commonly chosen as indicating failure, or alternatively a point approaching 100 per cent arcing may be so designated.

There are various methods for increasing the gradient available from a particular generator, such as decreasing electrode spacing, decreasing radius of curvature of an electrode, establishing a standing wave with a voltage maximum at the region of interest, or setting up a resonance condition with an electric field maximum at the region of interest [36a], [36b].

(e) Corona

Corona is a particular type of high-voltage discharge differing from arcing principally in the magnitude and

continuity of current involved. At low frequencies and high pressures, the difference is clearly evident: corona is a relatively low-current, often stable, discharge originating at a point of high gradient and diffusing into space; as the gradient is increased, there is a rapid transition to breakdown by arcing, which is a high-current cumulative discharge between two points, generally resulting in complete failure of the system during the arc. At high frequencies, however, electron oscillation within the region of high gradient can reduce the cumulative action of an arc to such a degree that the transition region between corona and arcing becomes entirely obscured, particularly at reduced pressures. For this reason, high-frequency breakdown is often considered to comprise all detectable discharges, including corona.

2.11 Measurement of Mode Purity

In recent years, especially for the higher frequencies, increasing use is being made of waveguides which may propagate several modes simultaneously. Generally, however, only one of these modes is of interest; the remainder are referred to as undesired or spurious modes. Two quantities are of interest in this connection: mode purity and spurious mode level. Mode purity generally refers to the desired mode and is defined as the ratio of power present in the forward-traveling wave of the desired mode to the total power present in the forward-traveling waves of all modes. Spurious mode level involves a ratio of the power present in the forward-traveling wave of a particular undesired mode to that in the desired mode; this ratio is commonly expressed in decibels.

(a) The amount of power present in the forward-traveling wave of any given mode is best measured by the use of a mode-selective device such as a coupled line transducer [37], [38]. In general, a separate device of this type is required for each mode of interest. Such a procedure enables one to completely determine the spurious mode level or to find the numerator of the mode-purity expression. The denominator of the mode-purity expression, the total power present in the forward-traveling wave of all modes, can be measured by any of the means indicated in Sec. 2.5 which are not mode-dependent; *e.g.*, a calorimetric method.

(b) The proportion of power in the forward-traveling waves of each of the propagating modes present may be measured directly by terminating the line in a mode-independent match and by appropriately probing the field distribution in the cross section. Such a probing procedure is rather difficult, however, and is feasible only for simple, stable field configurations and for a few modes present. Since at any given cross-section plane an arbitrary but constant time phase between the modes may exist, it is generally necessary to take measurements at more than one cross-section plane. In view of these difficulties, the probing is considered useful only for qualitative analyses.

(c) In the specific but valuable case of the TE_{01} mode in circular waveguide in which the TE_{02} mode is beyond cutoff, the radial resistive card mode filter [37] is sufficiently selective to permit the transmitted power to be regarded as being in the TE_{01} mode only and, therefore, to permit a direct measurement of the power in that mode.

(d) If the TE_{11} mode in a particular orientation in circular waveguide is designated as the desired mode, the orthogonal TE_{11} mode may be considered as a spurious mode. This latter mode may be excited by ellipticities or irregularities in the waveguide. The resulting polarization will, in general, be elliptical and may vary with location along the waveguide. Elliptical polarization may be completely described by the axial ratio and the orientation of the major axis of the ellipse of polarization. Mode purity may be determined at any particular cross section by noting the ratio of minimum to maximum field strength (axial ratio) and the angular position of the minimum with respect to the desired reference direction [39]. This measurement may be made by circumferential rotation of a radial detecting probe.

If the axial ratio is designated as $\tan \alpha$, and the angle between the direction of the maximum and the reference direction of the desired TE_{11} component is β, the level of the undesired TE_{11} component, $\tan^2 \gamma$, and the mode purity, $\cos^2 \gamma$, can be obtained from

$$\cos 2\gamma = \cos 2\alpha \cos 2\beta.$$

2.12 Measurement of Transmission Line Unbalance

(a) In the case of uniconductor waveguides or coaxial transmission lines, the question of unbalance does not usually rise. For a coaxial line, the total current on the inner conductor must always be equal and opposite to the total current on the inner surface of the outer conductor. The only balance problem which commonly occurs with a coaxial line is for the case when the generator or load is connected to the line in such a way that currents are caused to flow on the outer surface of the outer conductor. For this case, the presence of current on the outer conductor can be detected by a measurement of the magnetic field strength on the outer surface as outlined in Sec. 2.4.

(b) For a balanced transmission line system operating properly, the voltages (relative to ground) on the conductors are push-pull voltages, and the currents in the conductors are push-pull currents. When the system becomes unbalanced, this is evidenced by the presence of push-push voltages and currents in addition to the push-pull voltages and currents. The amount of the system unbalance is measured by the ratio of the push-push voltage component to the push-pull voltage component or by the ratio of the push-push current to the push-pull current (Sec. 2.1).

(c) One excellent method for measuring the ratio of unbalanced to balanced currents involves the use of a transmission line arrangement and shielded loop pickup

with a configuration such that in one position, the loop reads a value proportional to the unbalanced component of the current, while in a second position perpendicular to the first position, the loop reads a value proportional to the balanced component of the current [40]. This technique is one of the most accurate although it requires a careful set-up.

(d) A simple method for measuring the relative amount of unbalance involves the measurement of the voltage or the current on each conductor as a function of position [41] by one of the methods described in Secs. 2.1 or 2.2. If unbalance is present, then the standing wave on one conductor will be displaced from the standing wave on the second conductor. The amount of unbalance is related to the relative amount of displacement between the two standing waves. Unfortunately, this method will not give the absolute unbalance ratio since the voltage or current measurements give only the magnitude of either of these quantities. In order to determine the absolute ratio between the unbalanced and balanced quantities, it is necessary to know the complex values of either the voltage or the current in each of the two conductors at a given cross section.

Section 3—Measurement of One-Port and Two-Port Waveguide Components

It should be recalled that the performance of any waveguide component is affected by and must be specified with respect to the input and output waveguides between which it is inserted. The input and output ports need not be in the same kind of waveguide nor utilize the same mode of transmission, recalling, of course, that each port is identified with a single mode. As mentioned in Sec. 1, these measurement methods apply to linear, passive, reciprocal waveguide components.

3.1 Measurement of Insertion Loss [42]–[44]

According to the definition of insertion loss used in this Standard (see Sec. 5), both the load and the generator must be respectively matched to the waveguides which connect them to the waveguide component under test. It should be noted that this is not always the condition for maximum power transfer through the component. The insertion loss measured under these conditions is the sum of the dissipative and reflection losses.

Measurement of insertion loss can be made by noting the change in power level at a matched detector upon insertion of the component. An alternative method is a substitution method in which a matched calibrated variable attenuator is inserted in place of the component and adjusted to obtain the original power level at the detector. This attenuator may be part of the original measurement system, in which case the insertion loss is equal to the change in attenuation of the calibrated attenuator. When the input and output waveguide connections do not permit direct insertion of the component under test, it may be possible to make this measurement for high-loss components by including a low-loss adapter where convenient. The loss of the adapter should be small compared to the tolerable error of the measurement.

For very low-loss components, the method described in Sec. 3.9 is preferable.

3.2 Measurement of Input SWR

This measurement is made by measuring the standing-wave ratio (SWR) in the waveguide connected to the input port of the component, with the output port (if any) match-terminated. Any of the methods described in Sec. 2.9 (b) and (c) can be used. For lossless or symmetrical components, the result would be the same regardless of which port is used as the input.

3.3 Measurement of Input Impedance

This measurement is made by measuring the input impedance looking into one of the ports, with the other port (if any) match-terminated. Any of the methods described in Sec. 2.8 can be used. For components which are unsymmetrical, this impedance is not necessarily the same at both ports.

3.4 Measurement of Normalized Input Impedance

This measurement is made by measuring the normalized input impedance looking into one of the ports, with the other port (if any) match-terminated. Any of the methods described in Sec. 2.9 can be used. For unsymmetrical components, this impedance is not necessarily the same at both ports.

3.5 Measurement of Phase Shift

The phase shift through a waveguide component at a single frequency is the phase difference under matched conditions between corresponding incident and transmitted field quantities at the input and output ports, respectively, of the component, ignoring multiples of 2π radians. Since the phase difference will vary with load conditions, phase shift has been defined under matched conditions.

Phase shift measurements are based on the comparison of the phases of two signals by interference methods. The two signals may be obtained by sampling with directional devices at the input and output ports of the component under test [45], or they may be the signals transmitted through the waveguide containing the component under test and a reference waveguide of known phase shift [46], when equiphase signals are applied to the two waveguides. The comparisons may be made by employing hybrid junctions or by applying the two signals to opposite ends of a slotted section of waveguide [45] with suitable isolation to eliminate reflections. When a slotted section is employed in this manner, the position of a voltage minimum is observed with the component in the test apparatus. The phase shift of the component is then equal to twice the displacement of the minimum expressed in electrical degrees. Comparison procedures using a hybrid junction

employ null indications which require the use of a calibrated phase shifter in one of the arms [46]. A difference in the levels of the two signals being compared results in a finite minimum rather than a null indication. This effect does not affect the accuracy, but may influence the precision of the measurement.

The phase shift can also be determined by use of a sliding-short-circuit method to determine the appropriate component of the scattering matrix, as described in Sec. 3.9.

3.6 Measurement of Envelope Delay [47]

The quantity "envelope delay" (τ) in seconds is related to the rate of change of phase shift (ϕ) in radians with frequency (f) in cycles per second by the definition

$$\tau = \frac{1}{2\pi}\frac{d\phi}{df}.$$

For a nondispersive waveguide component, such as an air-filled TEM waveguide, the envelope delay is identical with the delay of the envelope of the wave.

The envelope of a wave passing through a dispersive waveguide component is distorted, so that the concept of the delay of the envelope loses meaning when the dispersion is high. While the delay of the distorted envelope cannot be strictly defined in a dispersive medium, to the degree to which the quantity is definable it is given by (τ).

The quantity $d\phi/df$ is obtained by measurements of the phase shift at several frequencies in the vicinity of the frequency at which the delay is to be evaluated. In measuring phase shift, the methods and restrictions of Sec. 3.5 are applicable.

The envelope delay may also be measured directly by the use of short pulses and oscillographic presentation of the time between input and output pulses.

3.7 Measurement of Power-Handling Capacity [33], [34], [48], [49]

The power-handling capacity of a waveguide component is limited by the occurrence of dielectric voltage breakdown, or by the effects of heating, or a combination of both. The energy in a single pulse may have to be limited to avoid overheating.

The limitation of power capacity by heating must be judged for each individual component. In some cases, radio frequency power cables for example, this limitation is measured by applying equivalent low-frequency power to the component.

The methods and restrictions applicable to the measurement of the dielectric voltage breakdown of a waveguide component are described under Sec. 2.10.

3.8 Measurement of Q [50], [51]

As here used, the Q of a resonant waveguide circuit is 2π times the ratio of the energy stored to the energy dissipated per cycle in that circuit. The following discussion applies to the Q at resonance of components exhibiting a simple resonance response and having a Q high enough (ordinarily about 10 or greater) to produce the desired accuracy of measurement.

If the circuit consists of a component under test, coupled to its associated loads, the resultant Q is termed the loaded $Q(Q_l)$ of the circuit, and is commonly referred to as the Q_l of the component under these conditions. For the component alone, without its coupling mechanisms, the pertinent quantity is the unloaded $Q(Q_u)$. These quantities are related by

$$\frac{1}{Q_l} = \frac{1}{Qu} + \frac{1}{Qe}$$

where Q_u results from dissipation in the component itself, and Q_e from dissipation in the external circuit through the coupling mechanism.

The Q of a component such as an inductor or a capacitor at any frequency is understood to mean the Q which would be obtained by resonating that component with a lossless element at that frequency.

Q_l may be measured directly, while Q_u is a derived quantity. The Q_l of a component depends on the particular coupling mechanisms; matched external loads are commonly used. In the case of a very loose coupling, the external losses become negligible, and Q_l approaches Q_u.

There are three basic methods of measuring the loaded $Q(Q_l)$. These are, first, the measurement of the transmission through the component as a function of frequency; second, the measurement of the time rate of decay of energy stored in the component at a given frequency, and third, the measurement of the input impedance of the component as a function of frequency.

(a) For the first method, it is necessary to determine the difference (Δf) between frequencies at which the power transmission differs by 3 decibels from the transmission at the resonance frequency, f_0. Then Q_l is obtained from the relation

$$Q_l = \frac{f_0}{\Delta f}.$$

If the component is symmetrically coupled to match-terminated lines of equal characteristic impedance, a measurement of its insertion loss (Sec. 3.1) at resonance will permit the unloaded $Q(Q_u)$ to be computed from the relation

$$\frac{1}{Q_u} = \frac{1-a}{Q_l},$$

where a = voltage transmission coefficient. The quantity a is related to the insertion loss (L) by

$$L = 20 \log_{10} (1/a) \text{ in decibels.}$$

It should be noted that when the insertion loss is low, this method may be inaccurate because of the difficulty in measuring the insertion loss to a sufficient degree of accuracy.

For example, if the resonator is a two-port component

which is designed for maximum transmission at resonance, the measurement may be made as follows. With a constant available power from the generator, the transmitted power is observed as a function of frequency in the vicinity of resonance, and the frequencies above and below resonance at which the transmitted power has been reduced by 3 decibels from the peak value at resonance are recorded. The transmitted power level may be determined by one of the methods described in Sec. 2.5. When the available power varies with frequency, the insertion loss of the component should be measured by one of the methods of Sec. 3.1 and 3-decibel change in insertion loss used to determine the appropriate frequency difference, Δf.

If the resonator is a two-port component designed for maximum absorption at resonance, the measurement procedure is similar to the above, except that the attenuation by the component is considered instead of transmission.

At low frequencies, a special case of this method is the measurement of bandwidth between frequencies where the voltage or current has fallen to 0.707 of its maximum value with a constant current or voltage source, respectively, and gives the unloaded $Q(Q_u)$.

(b) The second method of measurement, often referred to as a "decrement" measurement, is useful when the circuit has a sufficiently high Q so that it is simpler to make time interval measurements rather than frequency difference measurements.

It involves measuring the time interval (τ) required for the field strength at any point in the circuit to decay to $1/e$ (approximately 0.368) of its initial value after the generator output power has been suddenly interrupted. It is also necessary to measure the frequency (f) of the circuit resonance. The Q then follows from the relation

$$Q_l = \pi \tau f.$$

(c) The third of these methods, applicable to one-port components, is based on the relation

$$Q_u = \frac{\omega}{2G}\frac{dB}{d\omega} = \frac{\omega}{2G'}\frac{dB'}{d\omega} \qquad \text{at } \omega = \omega_0,$$

where G and B are the components of the input admittance seen at a plane at which $B=0$ at resonance, G' and B' are the components of the normalized input admittance at the same plane, and ω is the angular frequency. There exists a variety of specific methods for evaluating the necessary quantities, and the methods for measuring input admittance and normalized input admittance are those described in Secs. 2.8 and 2.9 herein.

3.9 Measurement of Scattering Matrix Elements by Sliding-Short-Circuit Methods [52]

3.9.1 General Discussion

The measurement of the scattering matrix elements or the equivalent circuit of a waveguide component by the sliding-short-circuit method is often the most accurate and convenient way of determining any or all of its properties. Sliding-short methods have the advantage of greater accuracy in that no relative power measurements are required, and no calibrated attenuators or carefully matched loads are needed. In the case of lossless components, only distance and frequency measurements are involved. These methods are also useful in measuring components through mismatched junctions.

The choice of scattering matrix or equivalent circuit representation depends on the application and the nature of the component to be measured. For example, if the component is to be employed with a matched load following it, particularly if only the transmission and reflection coefficients are of interest, the scattering matrix is clearly preferable. If the component is representable by a simple shunt or series element, use of the equivalent circuit is indicated. Even if the component does not permit a simple representation the pictorial features of an equivalent circuit may still be found useful, or a graphical procedure may be employed from which some pertinent partial information may be readily obtained. For all choices of representation the data is taken in basically the same fashion; the treatment of the data, however, depends on the representation desired. In this section the scattering matrix elements are considered. The equivalent circuit determinations are treated in Sec. 3.10.

3.9.2 Basic Measurement Procedure

The measurement set-up for the sliding-short method requires placing the component under test between a slotted section and sliding-short-circuit [53], [54]. The basic procedure consists of moving the sliding-short-circuit through a succession of known positions and determining the position of the voltage minimum (and VSWR if the component is dissipative) in the slotted section for each of the sliding-short-circuit positions.

3.9.3 Parameters Derived from the Measurement

The measured data, obtained in the manner indicated above, may be appropriately treated to yield the elements S_{11}, S_{12}, and S_{22} of the scattering matrix. These elements are complex, with magnitude and phase given by

$$S_{11} = |S_{11}|e^{j\phi_{11}}, \quad S_{12} = |S_{12}|e^{j\phi_{12}}, \quad S_{22} = |S_{22}|e^{j\phi_{22}}.$$

The magnitudes, but not the phases, of the elements are independent of input and output reference plane locations. The choice of reference plane locations is arbitrary; it may be based on physical convenience or computational simplicity. An example of the latter would be a choice such that ϕ_{11} and ϕ_{22} are both zero.

The scattering matrix elements have the following physical meaning: S_{11} is the input voltage reflection coefficient obtained when a matched load is placed at the output port; S_{22} is defined similarly to S_{11} except that

the component is reversed; S_{12} is the voltage transmission coefficient which is the ratio of the complex voltages of the wave transmitted past the component to that incident on it. Usually the $|S_{12}|^2$ is of particular interest, rather than the complex quantity S_{12}, since it is directly the power transmission coefficient of the component.

When the component is placed between a load and generator, each of which is matched to its waveguide, the magnitudes of the scattering matrix elements are simply related to the insertion loss (L_I), the reflection loss (L_R), and the dissipative loss (transmission loss, L_D) of the component. These relations are:

$$L_I = 10 \log_{10} \frac{1}{|S_{12}|^2}$$

$$L_R = 10 \log_{10} \frac{1}{1 - |S_{11}|^2}$$

$$L_D = 10 \log_{10} \frac{1 - |S_{11}|^2}{|S_{12}|^2}.$$

Another quantity of interest which can be derived from the above measurements is the minimum or intrinsic loss. This is the insertion loss (L_I) obtained when reactive elements producing maximum power transfer are placed at both ports of the two-port component. The resulting network, including these reactive elements, will be bilaterally matched ($S_{11} = S_{22} = 0$).

3.9.4 Dissipative Loss

The dissipative loss of a component may be obtained accurately by the sliding-short-circuit method without analyzing the data completely and obtaining the complete scattering matrix. The measurement procedure described under 3.9.2 above is followed; however, the procedure requires the power flow through the component to be opposite that for which the transmission characteristics are desired [55]. If the component possesses similar input and output connections, reversal of the component is all that is required. The data obtained are then plotted on the reflection coefficient chart. The radius (R) normalized to the chart radius of the resulting circular locus is then related to the dissipative loss, *i.e.*,

$$L_D = 10 \log_{10} \frac{1}{R}.$$

3.9.5 Complete Scattering Matrix

The complete scattering matrix for a component, which includes the phases of the above elements, may be determined by means of a method due to Deschamps [53], [56]–[58]. This method imposes an additional requirement on the measurement procedure outlined under 3.9.2 in that the short-circuit locations are most conveniently taken in pairs spaced a quarter guide wavelength apart. The resulting data are then plotted on the reflection coefficient chart, and the magnitudes and phases of the scattering matrix elements are separately obtained by means of graphical constructions.

3.9.6 Relation to Other Parameters

Some of the elements of the scattering matrix as determined by the method of 3.9.5 above are related simply to other parameters of the component as follows:

a) Input SWR $= \dfrac{1 + |S_{11}|}{1 - |S_{11}|}$

b) Normalized input impedance $= \dfrac{1 + S_{11}}{1 - S_{11}}$

c) Phase shift = angle of $S_{12} = \phi_{12}$.

3.9.7 Lossless Components

The scattering matrix of a lossless component is characterized by only three independent numbers, in contrast to six for a lossy component. The elements of the scattering matrix of a lossless component are related in the following fashion:

$$|S_{11}| = |S_{22}|,$$
$$|S_{12}|^2 = 1 - |S_{11}|^2,$$
$$2\phi_{12} = \phi_{11} + \phi_{22} \pm \pi.$$

In the measurement method referred to in 3.9.5 above, the data circle on the reflection coefficient chart becomes coincident with the unit circle, and the graphical constructions become somewhat simplified [53], [56]–[58].

An alternative measurement procedure for lossless components is the tangent relation method [53], [54], [58]. The measured data are taken in the manner described under 3.9.2 above. If the positions of the voltage nulls in the input and output waveguides relative to the chosen reference planes, T_1 and T_2, (measured away from the junction as shown in Fig. 1) are denoted by D and S, respectively, a simple plot of D vs S as indicated in Fig. 2 yields three real parameters D_0, S_0 and γ which characterize the lossless component. D_0 and S_0 are the values of D and S at the point of maximum slope, (γ). These parameters are related to the scattering matrix elements by:

$$|S_{11}| = |S_{22}| = \frac{\gamma + 1}{\gamma - 1}.$$

$$|S_{12}| = \frac{2\sqrt{-\gamma}}{1 - \gamma}$$

$$\phi_{11} = \frac{4\pi}{\lambda_g} D_0$$

$$\phi_{22} = \frac{4\pi}{\lambda_g} S_0 + \pi$$

$$\phi_{12} = \frac{2\pi}{\lambda_g} D_0 + S_0 + n\pi, \qquad (n = 0, 1).$$

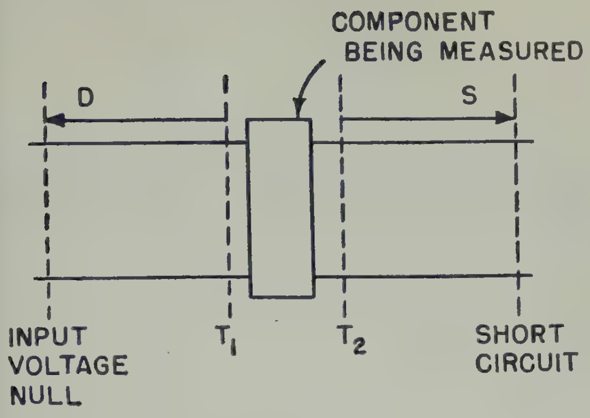

Fig. 1

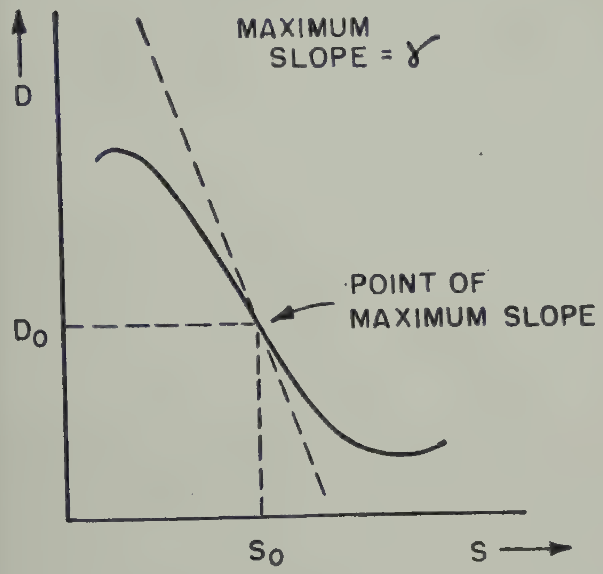

Fig. 2

Note: $(-\gamma)$ is a positive number and the sign of S_{12} is not determined.

3.9.8 Measurements Through a Junction

In certain cases it is necessary to perform measurements through a reflecting and possibly lossy junction or component such as an adapter. A modification of the graphical constructions described in 3.9.5 above may be employed to separate out the parameters of the component under test [57], [59], [60]. If the junction is lossless, an alternative and simpler procedure is available which employs the tangent relation method [61], [62]. Preliminary measurements of the junction alone are necessary in both methods.

3.9.9 Components with Small Loss

The measurement of the scattering parameters of a component with small dissipative loss is sometimes inaccurate or difficult because of the high values of SWR that need to be measured. These values of SWR can be lowered so that they fall into a more accurately measurable range by the deliberate addition of dissipative loss into the overall system. A simple additional measurement suffices to determine the effect of this added loss, which can then be subtracted out in order to obtain the properties of the component alone. A common and useful means for obtaining this added loss is a lossy shorting plunger [63].

3.10 Measurement of Equivalent Circuit Parameters by Sliding-Short-Circuit Methods [52], [53]

3.10.1 General Discussion

An equivalent circuit representation for an arbitrary dissipative component contains six real parameters (which appear as three complex parameters if the tee or pi network form is used). Symmetry in the component and in the location of the reference planes reduces this number to four; for the case of an unsymmetrical lossless component the number becomes three, while for a symmetrical lossless component only two independent real parameters are required. These parameters may be combined into any one of a variety of pictorial forms, in which the numerical values of the parameters vary with the reference plane location (since a different length of waveguide is absorbed into the equivalent circuit representation). It is sometimes convenient, by appropriate adjustment of the reference plane locations, to simplify the form of the equivalent circuit. The equivalent circuit for a lossless component can always be reduced to a shunt or series form or to a single ideal transformer by appropriate shifting of the reference plane locations; except in special cases, however, these final reference plane locations are not known in advance. Such reductions in form are also possible for lossy components; however, since only two reference plane shifts are available, the number of parameters for an arbitrary lossy component can be reduced only to four (or to three for a symmetric lossy component). The parameter values of an equivalent circuit are valid only at the frequency at which the measurement was made, and, in general, no inference may be drawn as to the frequency dependence of the parameters. This situation is similar to that occurring with the scattering matrix representation; however, for special cases, such as a very thin transverse iris in a uniconductor waveguide which is not operated near the cutoff frequency of the next mode, the frequency dependence may be inferred and this property of the equivalent circuit representation may be found useful.

3.10.2 Equivalent Circuits

A large variety of equivalent circuits may be found to correspond, at the same set of reference planes, to a given impedance matrix, of elements Z_{11}, Z_{22}, and Z_{12}. One of these equivalent circuits is the tee circuit of Fig. 3. In a similar manner, the pi circuit corresponds directly to the admittance matrix in Fig. 4. Another circuit representing the impedance matrix Z which is useful for lossless components is shown in Fig. 5 which contains an ideal transfomer.

$$n = Z_{11}/Z_{12}$$

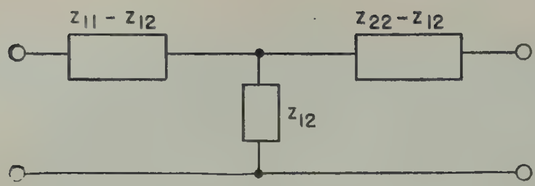

Fig. 3

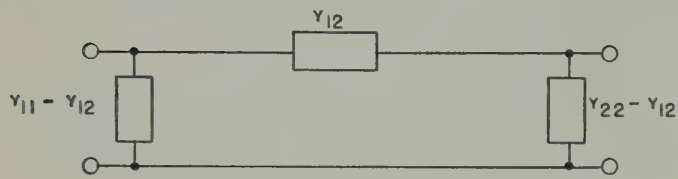

Fig. 4

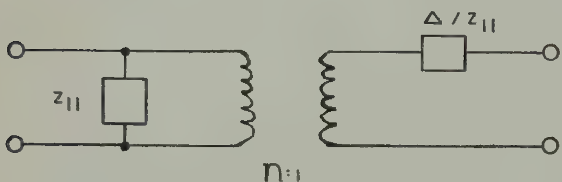

Fig. 5

where

$$\Delta = Z_{11}Z_{22} - (Z_{12})^2.$$

If the component is lossless, appropriate shifts of both the input and output reference planes reduce this network to one consisting of the shunt element only, or the transformer only [54]. The relations between the network parameters and the impedance matrix elements indicate readily that to eliminate the series element, Δ must be zero, or that to eliminate the transformer, Z_{11} must equal Z_{12}. A dual circuit corresponds to the admittance matrix.

3.10.3 Shunt or Series Networks

When the component under test can be represented by a purely shunt or series equivalent circuit at particular reference planes, the parameters may be determined from a single input impedance measurement. For a series network, the normalized resistance and reactance are given directly by the normalized input impedance to the component when a short circuit is placed at the output reference plane. For a shunt network, the normalized conductance and susceptance follow directly from the measurement of the normalized input admittance to the component when an open circuit is placed at the output reference plane.

3.10.4 Simplified Method for Symmetrical Components

When the component is symmetrical but not representable by a purely shunt or series equivalent circuit, the parameters may be determined by measuring the input impedances at the input reference plane corresponding to both a short and an open circuit termination at the output reference plane. The normalized impedance matrix elements Z_{11}', Z_{22}', and Z_{12}' are related to the measured quantities by [53]:

$$Z_{11}' = Z_{22}' = Z_{oc}'$$

$$(Z_{12}')^2 = Z_{oc}'(Z_{oc}' - Z_{sc}')$$

$$Z_{11}'Z_{22}' - (Z_{12}')^2 = Z_{sc}'Z_{oc}'$$

where Z_{oc}' and Z_{sc}' are the measured normalized input impedances at the input reference plane corresponding to an open circuit and a short circuit, respectively, at the output reference plane. Note that the sign of Z_{12}' is not determined by these relations. The above relations are also valid for the admittance matrix elements if all the normalized impedances are replaced by normalized admittances, and if the role of open and short circuits is reversed.

Note: If the component is unsymmetrical, the above procedure can be generalized by requiring an additional terminating impedance and its corresponding input impedance measurement. However, the relations then become more involved, and the usefulness of the procedure becomes questionable.

3.10.5 Lossless Components

For lossless components, the tangent relation method furnishes an alternative procedure to the preceding ones in Sec. 3.10. This method is more accurate because additional points are taken and the data are averaged in a systematic fashion. The measurement procedure, which employs a sliding short circuit, is the same as that described under 3.9.2. If, then, the positions of the voltage nulls in the input and output waveguides relative to the chosen reference planes are denoted by D and S, respectively, a simple plot of D vs S (see figures associated with 3.9.7) yields three real parameters D_0, S_0, and γ which characterize the lossless component [53], [54], [58]. These parameters (defined in Sec. 3.9) are related to the normalized reactance matrix elements by:

$$\frac{X_{11}}{X_{01}} = -\frac{\alpha\beta + \gamma}{\beta - \alpha\gamma},$$

$$\frac{X_{22}}{Z_{02}} = \frac{1 + \alpha\beta\gamma}{\beta - \alpha\gamma}$$

$$\frac{X_{11}X_{22} - (X_{12})^2}{Z_{01}Z_{02}} = -\frac{\alpha - \beta\gamma}{\beta - \alpha\gamma},$$

where

$$\alpha \equiv \tan \frac{2\pi}{\lambda_{g1}} D_0,$$

$$\beta \equiv \tan \frac{2\pi}{\lambda_{g2}} S_0,$$

and where the subscripts 1 and 2 refer to the input and output waveguides, respectively. The above relations are also valid for the elements of the susceptance matrix if X is replaced everywhere by B, and Z by Y, and if

$$\alpha = -\cot \frac{2\pi}{\lambda_{g1}} D_0$$

$$\beta = -\cot \frac{2\pi}{\lambda_{g2}} S_0.$$

These relations are also useful for determining the appropriate reference plane shifts required for obtaining simplified equivalent circuit forms.

3.10.6 Lossy Components

Methods also exist for lossy components in which the equivalent circuit parameters are obtained by sliding-short methods. One of these methods is applicable to unsymmetrical lossy components, and yields an equivalent circuit which is almost shunt or almost series, and in which the lossy and lossless portions are separated from each other [53], [58], [64]. A second method, suited to symmetrical lossy components, is based on the representation of the lossy portion by an ideal attenuator [65], [66]. For the same degree of accuracy, the effort involved in the analysis of the measured data to yield the final equivalent circuit in these methods is about the same as that required in the procedure discussed in 3.9.5 for determining the complete scattering matrix.

Section 4—Measurements of Multi-Port Waveguide Components [67], [68]

Multi-port components are herein considered as those having more than two ports. The ports need not be in the same kind of waveguide nor utilize the same mode of transmission, recalling that each port shall be identified with only a single mode. Again, only linear, passive, reciprocal networks are treated.

In general, the measurements of multi-port components reduce to the measurements described in Secs. 3.1 through 3.8 with a specified pair of ports comprising the input and output with the remainder of the ports match-terminated. In addition, certain properties of multi-port junctions consist of a comparison of the transmission properties of specific pairs of ports. It should be noted that the reflection of a specified port and transmission of a port pair may be substantially affected by the load conditions at the other ports.

Terms commonly used to describe the coefficient of transmission from one port to another in a multi-port waveguide component are "coupling" and "isolation." Isolation (sometimes called "decoupling") is identical with insertion loss, whereas coupling is the negative of insertion loss, expressed in decibels. The use of these terms depends on the intent of the application; coupling usually refers to a definite desired transmission coefficient between a particular pair of ports, whereas isolation refers to a transmission coefficient which should be as small as possible. The term "unbalance" may be used to refer to the difference between the insertion losses between an input port and each of two output ports where these are desired to be as nearly equal as possible.

4.1 Measurement of Coupling [69], [70]

The coupling between any two ports of a multi-port component is specified by the insertion loss between these ports and is measured by the methods of Sec. 3.1. The numerical value attached to the term "coupling" is the negative of the insertion loss when expressed in decibels or, equivalently, the reciprocal when expressed as a power ratio.

A common example of a four-port component is the symmetrical directional coupler represented by Fig. 6. For this component, the negative (or reciprocal) insertion loss from port 1 to port 4, or from port 2 to port 3, would be termed the forward coupling, or simply the "coupling" of the directional coupler. These couplings are related to the elements S_{41} and S_{32}, respectively, of the scattering matrix of the component. For an unsymmetrical directional coupler, the elements S_{41} and S_{32} may be different, and the term coupling should be associated with the direction of use.

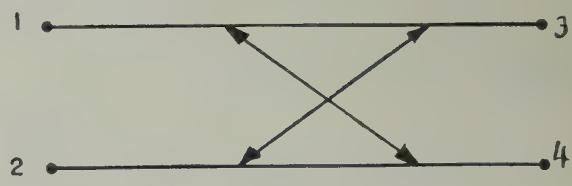

Fig. 6—Symmetrical directional coupler.

4.2 Measurement of Isolation

The isolation between any two ports of a multi-port component is given directly by the insertion loss between these ports when the other ports are match-terminated. The accuracy of the isolation measurement may be critical to departures from matched load conditions on the other ports of the component.

Special techniques are available to correct for the inaccuracies introduced by the loads [71]. The measurement is generally made with a sensitive detector circuit.

A typical four-port component for which the measurement of isolation is significant is the hybrid tee shown in Fig. 7 [71]. The isolation between the E and H arms of this component is the insertion loss between ports 1 and 2: the isolation between collinear arms is the insertion loss between ports 3 and 4. These isolations, which are not necessarily equal, are related to the scattering matrix elements S_{21} and S_{43}, respectively.

The insertion loss between ports 1 and 2 or ports 3 and 4 of the symmetrical directional coupler of Fig. 6 is the isolation of the component; its negative (in decibels) is also frequently called the "backward coupling" of the coupler.

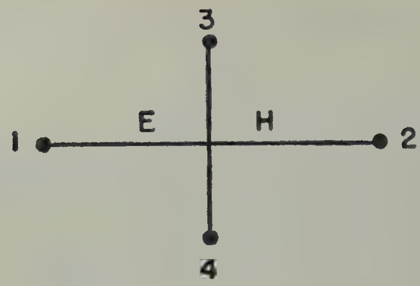

Fig. 7—Hybrid tee.

4.3 Measurement of Unbalance [72]

Small inequalities between the supposedly equal transmissions (magnitude and phase) from a common port to other ports, in turn, of a multi-port component are of interest in components such as simple tees, hybrid tees, and *n*-port power dividers. The power unbalance is the difference, in decibels, between the insertion losses between an input port and each of two output ports.

The outputs may be measured simultaneously or in sequence by matched detectors, or the difference in outputs may be observed with an appropriate amplitude or phase comparison circuit. In the case of simultaneous measurement, the accuracy may be improved by interchanging detectors, and averaging.

An alternative method for measuring unbalance consists of observing the small output from the input port when the output pair of ports are simultaneously fed equal magnitude signals of opposite phase.

4.4 Measurement of Directivity [73]

In the case of a directional coupler, the difference in insertion loss from one port of one waveguide to each of the two ports of the other waveguide is known as the directivity. In Fig. 6, for power fed into port 1, directivity is the insertion loss between ports 1 and 2 less the insertion loss between ports 1 and 4, with the insertion loss expressed in decibels.

Section 5—Definitions

Attentuation Constant. Of a traveling plane wave at a given frequency, relative rate of decrease of amplitude of a field component (or of voltage or current) in direction of propagation in nepers per unit length.

Characteristic Impedance (of a Two-Conductor Transmission Line). For a traveling transverse electromagnetic wave, the ratio of the complex voltage between the conductors to the complex current on the conductors in the same transverse plane with the sign so chosen that the real part is positive.

Electric (Magnetic) Field Strength. The magnitude of of the electric (magnetic) field vector.

Input Impedance of a Transmission Line. The impedance between the input terminals with the generator disconnected.

Insertion Loss (of a Waveguide Component). The change in load power, due to the insertion of a waveguide component at some point in a transmission system, where the specified input and output waveguides connected to the component are reflectionless looking in both directions from the component (match-terminated). This change in load power is expressed as a ratio, usually in decibels, of the power received at the load before insertion of the waveguide component to the power received at the load after insertion.

Note 1: A more general definition of insertion loss does not specify match-terminated connecting waveguides, in which case the insertion loss would vary with the load and generator impedances. In this Standard, match-terminated connecting waveguides will be assumed unless otherwise specified.

Note 2: When the input and output waveguides connected to the component are not alike or do not operate in the same mode, the change in load power is determined relative to an ideal reflectionless and lossless transition between the input and output waveguides.

Matched Termination (for a Waveguide). A termination producing no reflected wave at any transverse section of the waveguide.

Matched Transmission Line. See Matched Waveguide.

Matched Waveguide. A waveguide having no reflected wave at any transverse section.

Mode of Propagation (Transmission). A form of propagation of guided waves that is characterized by a particular field pattern in a plane transverse to the direction of propagation, which field pattern is independent of position along the axis of the waveguide.

Note: In the case of uniconductor waveguides the field pattern of a particular mode of propagation is also independent of frequency.

Mode Purity. The ratio of power present in the forward-traveling wave of a desired mode to the total power present in the forward-traveling waves of all modes.

Normalized Impedance (Relative to a Given Waveguide). An impedance divided by the characteristic impedance of the waveguide.

Note: The relation between voltage, current and power chosen for the characteristic impedance must also be taken for the impedance to be normalized, in which case the normalized impedance will be independent of the convention used to define the characteristic impedance.

Port (for a Waveguide Component). A means of access characterized by a specified reference plane and a specified propagating mode in a waveguide which permits power to be coupled into or out of a waveguide component.

Note 1: At low frequencies the port is synonymous with a terminal pair.

Note 2: To each propagating mode at a specified reference plane there corresponds a distinct port.

Reflection Coefficient (in a Transmission Medium). At a given frequency, at a given point, and for a given mode of transmission, the ratio of some quantity associated with the reflected wave to the corresponding quantity in the incident wave.

Note: The reflection coefficient may be different for different associated quantities, and the chosen quantity must be specified. The "voltage reflection coefficient" is most commonly used and is defined as the ratio of the complex electric field strength (or voltage) of the reflected wave to that of the incident wave.

Reflection Coefficient (of a Transition or Discontinuity). For a transition or discontinuity between two transmission media, the reflection coefficient at a specified point in one medium which would be observed if the other medium were match-terminated.

Scattering Matrix. A square array of complex numbers consisting of the transmission and reflection coefficients of a waveguide component.

As most commonly used, each of these coefficients relates the complex electric field strength (or voltage) of a reflected or transmitted wave to that of an incident wave. The subscripts of a typical coefficient S_{ij} refer to the output and input ports related by the coefficient. These coefficients, which may vary with frequency, apply at a specified set of input and output reference planes.

Standing Wave Ratio. At a given frequency in a uniform waveguide, the ratio of the maximum to the minimum amplitudes of corresponding components of the field (or the voltage or current) along the waveguide in the direction of propagation.

Note: Alternatively, the standing wave ratio may be expressed as the reciprocal of the ratio defined above.

Transmission Coefficient (of a Transition or Discontinuity). For a transition or discontinuity between two transmission media, at a given frequency, the ratio of some quantity associated with the transmitted wave at a specified point in the second medium to the same quantity associated with the incident wave at a specified point in the first medium, the second medium being match-terminated.

Transmission Line. A waveguide consisting of two or more conductors.

Transmission Loss. In the transmission of power past two points, the ratio, usually expressed in decibels, of the net power passing the first point to the net power passing the second.

Uniconductor Waveguide. A waveguide consisting of a cylindrical metallic surface surrounding a homogeneous dielectric medium.

Note: Common cross-sectional shapes are rectangular and circular.

Uniform Waveguide. A waveguide in which the physical and electrical characteristics do not change with distance along the axis of the guide.

Waveguide. A system of material boundaries capable of guiding waves.

Waveguide Component. A device designed to be connected at specified ports in a waveguide system.

Waveguide Wavelength. For a traveling wave in a uniform waveguide at a given frequency and for a given mode, the distance along the guide between similar points at which a field component (or the voltage or current) differs in phase by 2π radians.

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Correspondence

Stereo Frequency Response*

These comments are stimulated by those submitted by Sobel[1] who suggested that the frequency response necessary for stereo reproduction might be considerably curtailed from the maximum frequency response to which the ear responds. Sobel proposed that work might be done which could result in such a determination.

Extensive work has been done in this respect by the two foremost acoustical experts in the world. It has definitely been proven that the full frequency range the ear accommodates is not only preferred by comparative listener tests, but is necessary from the standpoint of an analysis of the separate effects of the binaural and stereophonic phenomena. This work has been described by Dr. Harry F. Olson, Director, Acoustical and Electromatical Mechanical Research Laboratory, RCA Laboratories, Princeton, N. J., and Dr. Harvey Fletcher, Director of Scientific Research, Brigham Young University, Provo, Utah, formerly Director of Physical Research, Bell Telephone Laboratories.

An account of Dr. Olson s work, concerning the listener preference tests, appears in his book.[2] In section 12.29, he describes extensive listening tests (after Chin and Eisenberg) which indicated that the listeners preferred a curtailed range of frequencies for reproduced monaural music. In section 12.30, he points out that the reason for this choice by the listeners was undoubtedly due to distortions in the reproduction equipment, and goes on to describe listener tests which employed live music and acoustical filters which resulted in a preference of the listeners for the full range for the case of both music and speech. The obvious result of these extensive tests is that when the reproduction is natural enough, the listener prefers the full range.

Similar frequency preference tests were conducted by Dr. Olson using stereophonic sound reproduction and are described in section 12.31. The subjective tests of frequency range preference were conducted for speech and music reproduced in auditory perspective employing a two-channel stereophonic sound system. The full frequency range was a flat response to 15,000 cycles. "The comparison range was a flat response to 5000 cycles. The results of these tests indicate a preference for a full frequency range" (for music). "The frequency preference for speech

* Received by the IRE, January 23, 1959. The author submitted this letter for comments to both Dr. Olson and Dr. Fletcher, and they agreed that the material is factual and correct.

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also indicates a preference for the full frequency range."

The work of Dr. Fletcher concerns the analysis of the separate stereo and binaural effects. In chapters 12 and 13 of his book[3] he points out that the binaural and stereo effects are dependent upon three factors: the *phase* effect, the *amplitude* effect, and the *sound quality*. A consideration of any one of these alone does not give the full effect. Many measurements have been taken considering the phase effect alone with blindfolded listeners who indicated the location of the image as the phase was varied while the amplitude was held constant. These measurements indicated that the phase effect becomes uncertain above approximately 1000 cycles. This reference, which considers only one of the three pertinent factors, may be responsible for Sobel's assumption that only a limited frequency response might be required to supply the full stereo effect.

In Chapter 13, "Auditory Perspective," Dr. Fletcher considers stereophonic transmission where two or more loudspeakers are employed. Much data are given which show the intricate nature of the amplitude perception of the stereophonic effect by the ear. A graph reproduced on page 225 shows that the amplitude perception of the stereophonic effect is greatest in the range of frequencies between 5000 and 15,000 cycles. He concludes that: "Of the factors influencing angular localization, loudness differences of direct sound seem to play the most important part; for certain observing positions the effect can be predicted reasonably well from computations."

On page 228 this statement is made: "If the quality from the various loudspeakers differs, the quality of sound is important to localization. In general, localization tends towards the channel giving the most natural or 'close-up' reproduction, and this effect can be used to aid the loudness differences in producing angular localization." This is an interesting point. It shows that AM-FM stereo, or the FM multiplex stereo of the type which employs full fidelity on the main channel of the FM transmitter and a curtailed fidelity on the multiplex sub-carrier channel, result in an improper localization. An interesting additional fact in this respect is the experiment described on page 216 in which different frequency ranges of speech and music were applied to the two channels. In the case of speech, "... the brain was able to combine the sounds obtained from the two ears to complete the proper picture. However, when music was transmitted a different situation resulted. This was particularly true when listening to music from the piano. In this case the tones appeared first in one ear and then in the other ear depending upon the pitch. This causes confusion and gives a very weird sort of sensation." The obvious conclusion of these extensive measurements is that while curtailed frequency response on one of the loudspeaker systems may give an apparent stereo effect, the full and most desirable effect cannot be obtained. A further conclusion is that with the compatible[4] system of FM multiplex stereo, which balances the frequency response of the two loudspeakers, the degradation produced by unequal frequency response is removed.

It is my feeling that we are in a formative stage with respect to stereo. Much is known about it, as is evidenced by the published material mentioned above. However, as this system is brought more and more into practice by the stereo tapes, disks, and broadcasting systems, still more will be learned to enhance it further. Hence, nothing should be done at this time which would place any limitation on such further developments. In addition, it appears that stereo is providing a new tool for the maestro. Interesting effects of separation are being employed by the stereo disk manufacturers such as very realistic soft-shoe dancing which dances from one loudspeaker to the other and gives a further imaginative effect of the listener being "there." Such devices as the two-studio technique which uses one microphone, in one studio with one group of musicians, another microphone in another studio with another group of musicians, and glass panels that permit both to be seen by the maestro, may appear as stunts with a certain amount of deceit. However, out of such stunts there will inevitably emerge new techniques which will further enhance musical reproduction for the greater enjoyment of the listening public. A free hand should be allowed in every respect. The listeners will judge which is good and which is bad, but they must have the opportunity for the best frequency response, distortion characteristics, and other factors, to make such judgment.

Sobel is to be commended for bringing the subject up as he did. I hope that these comments provide a clarification.

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[3] H. Fletcher, "Speech and Hearing in Communication," D. Van Nostrand Co., Inc., New York, N. Y.; 1953.

[4] M. G. Crosby, "A compatible system of stereo transmission by FM multiplex," *J. Audio Eng. Soc.*, vol. 6, pp. 70–73; April, 1958.

Directional Bridge Parametric Amplifier*

Mr. S. H. Autler in his letter[1] describes a proposed maser amplifier that does not require non-reciprocal elements. A similar system utilizing varactor diodes has been used at the Bell Laboratories.

The operation of this system can be seen from Fig. 1. A signal in arm 1 of the hybrid divides equally between arms 2 and 3. This signal is amplified by the varactors and reflected back to the hybrid. When the phase adjustment of the varactors and phase shifters shown on the figure in arms 2 and 3 are such that the returned signals at the hybrid are 180° out of phase, we have addition of the signals in arm 4 and cancellation in arm 1.

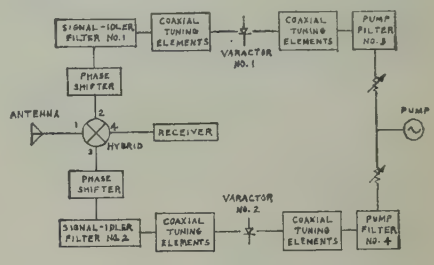

Fig. 1—Directional bridge parametric amplifier.

The noise figure of this directional bridge amplifier for a matched antenna and ideal balanced hybrid is the same as the single varactor. Heffner and Wade[2] have shown that the noise figure of a varactor is given by

$$F = 1 + \frac{G_1}{Gg} + \frac{G}{Gg}\frac{\omega_1}{\omega_2} + \text{higher terms.}$$

Normally,

$$\frac{G_1}{Gg} \simeq 0 \quad \text{and} \quad \frac{G}{Gg} \simeq 1.$$

hence the noise figure of the varactor in the degenerate mode (signal frequency approximately equal to idler frequency) is

$$F \simeq 1 + \left[\frac{\omega_1}{\omega_2}\right] \simeq 3 \text{ db.}$$

Of course, a mismatched antenna or unbalanced hybrid will cause an increase in noise figure above the 3-db figure.

The experimental circuit follows Fig. 1 closely. A broad-band coaxial hybrid was used. A varactor diode with its associated tuning elements was placed in each side arm (2 and 3). Means of controlling phase shift were placed between the filters and hybrid. Filters 1 and 2 pass the signal and idler frequencies but reject the pump. Filters 3 and 4 pass the pump but reject the signal and idler. The pump is fed from a common source through separate attenuators to the diodes. DC bias was supplied separately to the diodes.

The experimental results of the directional bridge amplifier operating at 530 mc (pump at 1060 mc) are promising. Gains up to 25 db have been obtained although the margin of stability was small. For a 10-db gain, the following results were obtained. A pump power of about 4 milliwatts was required. A bandwidth of approximately .6 mc was obtained. The hybrid balance at 530 mc was such that the VSWR in the input arm 1 was $1.02 \pm .01$. The average of the measured noise figures was 3.8 db which compares favorably with the theoretical noise figure of 3 db for this mode of operation.

* Received by the IRE, January 8, 1959.

[1] S. H. Autler, "Proposal for a maser-amplifier system without nonreciprocal elements," PROC. IRE vol. 46, pp. 1880–1881; November, 1958.

[2] H. Heffner and G. Wade, "Gain Bandwidth and Noise Characteristics of the Variable Parameter Amplifier," Stanford Electronics Lab., Stanford University, Stanford, Calif., Tech. Rep. No. 28.

Further experiments indicate that the directional bridge parametric amplifier can be used as a tunable amplifier by varying pump frequency, tuning, and phase shift. The range of operation would be limited by the broad-band nature of the hybrid. This directional bridge amplifier can be used at any frequency where a hybrid and a maser or reactance device are available.

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Minimum Insertion Loss Filters*

The recent letter by La Rosa[1] can be used, together with the papers referenced in the letter, to formulate a general problem whose solution has not been obtained yet: "Given the requirements for a band-pass filter in terms of bandwidths, selectivity, and such other requirements as particular problems suggest, chose the shape of the response and the design criteria that minimize the insertion loss in the center of the pass bands."

La Rosa shows that in a particular case a symmetrical filter meets the minimum loss conditions. On the other hand, Dishal has shown that, for the case of Butterworth or Tchebycheff shapes, asymmetrical filters are required to minimize the mid-band loss. Despite the apparent contradiction, both answers, of course, are correct and show that the general solution depends upon the initial requirements.

The importance of the problem is great enough to warrant effort by other researchers, and the purpose of this note is to formulate the general problem and invite attention to it.

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* Received by the IRE, November 13, 1958.
[1] La Rosa, "Optimum coupled-resonator band-pass filter," PROC. IRE, vol. 47, pp. 329–330; February, 1959.

AGU Committee for the Study of the Metric System in the United States*

Pursuant to the resolution adopted unanimously at the business session of the American Geophysical Union on May 7, 1958,[1] President Ewing appointed a Committee on the study of the Metric System in the United States.

METRIC SYSTEM QUESTIONNAIRE

1. Indicate professional field of interest in the AGU. ______
2. What approximate percentages of units used in your work are
   Metric______ British______
   Other______
3. Would it be to your advantage if a complete conversion to the metric system could eventually be made?
   Yes______ No______
4. How long a period of time should be allowed for the conversion, in years?
   10____ 20____ 30____ 40____ 50____
   Longer____
5. Should the centigrade system of temperature measurement be adopted?
   Yes______ No______
6. Do you believe that U. S. export trade is suffering due to the use of British units?
   Yes______ No______ No Opinion______
7. Do you believe that the eventual adoption of the metric system is inevitable?
   Yes______ No______
8. In the event of a ong time conversion to the metric system, do you believe that the cost would be proh bitive?
   Yes______ No______
9. In the event a Joint Committee were established to study the problem, circulate questionnaires, accumulate statistics, and report, it should be sponsored by (check one):
   Professional societies ______
   Educational institutions______
   Industry ______
   Government ______

   How should the study be financed?
   Would you be willing to assist such a study group?
   Financially ______
   As an advisor______
10. Additional remarks at any length are welcomed.

Bills for the exclusive adoption of the metric system in the United States have been presented to Congress more than once, but they failed, the principal reason being that the effective date proposed was entirely too soon after passage of the bill. An early effective date would undoubtedly work a severe hardship on the adult population not familiar with the metric system, and it would make obsolete a prohibitive amount of everyday items of weights and measures. A solution would appear to be a bill to make the metric system the only official system of weights and measures in the United States, effective in not less than one generation, 33 years, after passage of the bill. Following this action by the Congress, the grade schools and high schools would begin immediately to teach children the metric along with the English system and, during the transition period, would place more and more emphasis on the metric system. By the end of the transition period, the English system would still be taught, but the emphasis would be completely reversed from what it is today. In a generation, most items of equipment involving weights and measures normally become obsolete or worn out and are replaced. Also, persons engaged in professions and trades now using the English system exclusively would normally pass on to retirement during this period and would be replaced by a new generation thoroughly educated and trained in the metric system. A long transition period should result in a smooth change to this simplified decimal system under which 90 per cent of the world s people now live.

QUESTIONNAIRE

The accompanying questionnaire is directed to readers for the purpose of gathering statistical information to indicate the degree of interest in this matter. The metric committee of the American Geophysical Union will welcome any comments. Those submitting replies are urged to suggest solutions to difficulties which may be foreseen in the adoption of the metric system. Signature of these replies is optional.

Additional copies of the questionnaire are available upon request. A small effort on your part to complete and mail this questionnaire will be of invaluable help to the Committee. Address all correspondence to:

The Executive Secretary
American Geophysical Union
1515 Massachusetts Avenue N. W.
Washington 5, D. C.

FLOYD W. HOUGH
Committee Chairman

* Received by the IRE, January 30, 1959.
[1] *AGU Trans.*, p. 558; June, 1958.

Low Noise Parametric Amplifier*

In this note the authors report preliminary analytical and experimental results obtained with a cavity-type parametric amplifier at S-band in which through variable coupling the effect of diode losses on noise figure can be minimized at the expense of pump power. In this way, excess noise temperatures of 100°K have been obtained at room temperature. By cooling the diode with liquid nitrogen temperature, the excess noise temperature was reduced to 50°K.

In what follows it is assumed that the reader is acquainted with the referenced literature.[1,2,3] The noisiness of the amplifier is described by its "excess noise temperature T_e"[4] which is related to the "noise figure F"[5] as follows:

$$\frac{T_e}{T_0} = F - 1 \tag{1}$$

where

$$T_e = \frac{N_e}{kBG}$$

* Received by the IRE, January 19, 1959.
[1] J. M. Manley and H. E. Rowe, "Some general properties of nonlinear elements," Part II, PROC. IRE, vol. 46, pp. 850–860; May, 1958.
[2] S. Bloom and K. K. N. Chang, "Theory of parametric amplification using nonlinear reactances," *RCA Rev.*, vol. 18, pp. 578–596; December, 1957.
[3] H. Heffner and G. Wade, "Gain, band width and noise characteristics of the variable parameter amplifier," *J. Appl. Phys.*, vol. 29, pp. 1321–1331; September, 1958.
[4] J. C. Helmer and M. W. Müller, "Calculation and measurement of the noise figure of a maser amplifier," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-6, pp. 210–215; April, 1958.
[5] H. A. Haus and R. B. Adler, "Optimum noise performance of linear amplifiers," PROC. IRE, vol. 46, pp. 1517–1533; August, 1958.

N_e = excess output noise power generated between input and output terminals of the device
G = power gain
B = effective bandwidth $= 1/G^{\max}\int_0^\infty G(f)df$
T_0 = reference temperature (290°K)

Dur analysis is based on the equivalent circuit of Fig. 1. A single cavity supports both signal and idler frequencies and an ideal circulator is symbolically indicated by arrows showing direction of unattenuated powerflow. The subscripts g, d, p refer to generator, signal, diode, and pump respectively. The coupling coefficients k are defined as follows:

$$k_{sd} = Q_s/Q_{sd}, \text{ etc. } k_{sg} = Q_s/Q_{sg}, \text{ etc.}$$

where

Q_s is the unloaded signal circuit Q
Q_{sd} is the external signal circuit Q loaded by the diode
Q_{sg} is the external signal circuit Q loaded by the generator.

In addition, we define the following quantity:

$$\mu_g = Q_{sd}/Q_{sg}.$$

This coefficient measures directly the degree of coupling between the generator and diode. Extension of the above model to include a separate idler cavity is obvious. The gain and bandwidth expressions for this model are, of course, identical with the published ones provided the admittances and Q's are properly interpreted. The excess noise temperature and the critical pump power (power necessary for onset of oscillations) can be given for the general case in terms of the above quantities by (2) and (3).

$$T_e = T_s\frac{1}{k_{sg}} + T_d\frac{1}{\mu_g} + \alpha\frac{\omega_s}{\omega_i}\frac{k_{id}}{k_{id}+1} \cdot\left(1+\frac{1}{k_{sg}}+\frac{1}{\mu_g}\right)\left(T_d + T_i\frac{1}{k_{id}}\right) \quad (2)$$

$$P_{CR} = \frac{\omega_s C}{Q^3{}_d}\left(\frac{C}{\mathfrak{C}}\right)^2\left(1+\mu_g+\frac{1}{k_{sd}}\right) \left(1+\frac{1}{k_{id}}\right)\left(1+\frac{1}{k_{pd}}\right)\left(\frac{\omega_i}{\omega_s}\right)\left(\frac{\omega_p}{\omega_s}\right)^2 \quad (3)$$

where

$C_d = C + \mathfrak{C}V$ defines diode nonlinearity
Q_d = diode Q at signal frequency
T_s = temperature of signal circuit
T_i = temperature of idler circuit
T_d = diode temperature
T_g = temperature of generator resistance.

For the single cavity experiment in which the noise performance is measured with a broadband noise source, the above expressions reduce to (4) and (5).

$$\frac{T_e}{T_0} = \frac{T_s}{T_0}\frac{1}{k_{sg}} + \frac{T_d}{T_0}\frac{1}{\mu_g} \quad (4)$$

$$P_{CR} = 4\frac{\omega_s C}{Q_d^3}\left(\frac{C}{\mathfrak{C}}\right)^2\left(1+\mu_g+\frac{1}{k_{sd}}\right) \cdot\left(1+\mu_g+\frac{1}{k_{sd}}\right)\left(1+\frac{1}{k_{pd}}\right). \quad (5)$$

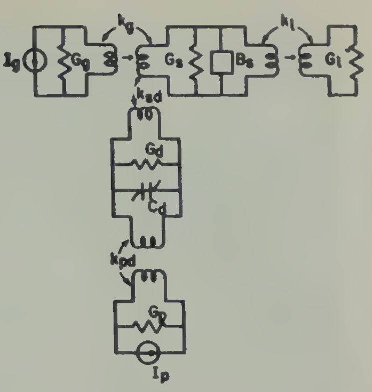

Fig. 1—Equivalent circuit of single-cavity parametric amplifier using ideal circulator.

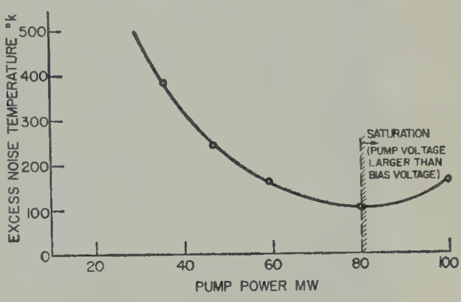

Fig. 2.

Eqs. (4) and (5) show the important characteristic that the amplifier's noise temperature can be reduced at the expense of increased pump power. This is obtained by increasing μ_g, *i.e.*, reducing the diode-to-signal circuit coupling relative to the generator-to-signal circuit coupling.

From (4) and (5) both noise figure and pump power for a specific amplifier can be rather accurately predicted because the diode parameters as well as the relative coupling coefficients (ratios of Q's) can be measured or evaluated. A verification of these relations was performed with an S-band (3100 mc) waveguide cavity amplifier with pump power equal to about twice the signal frequency. This amplifier was so designed that the diode coupling to the signal cavity could be varied. Provision was also made for cooling the diode with liquid nitrogen temperature without cooling the circuits. Noise figure was measured both with an argon discharge lamp and, in an absolute measurement, by reference to the noise from a matched load cooled to liquid nitrogen temperature. In these measurements, because idler and signal frequencies were about the same, noise from the reference noise source was fed to both signal and idler channels. This means that there was no contribution from the idler channel to the excess noise temperature and that (4) did apply.

The diodes used in these experiments were gold bonded diodes produced expressly for this purpose by the Semiconductor Division of Hughes Products. At this stage of development, the diode Q at 3000 mc and 2 volt reverse bias was about 15.

A typical variation of amplifier noise temperature with pump power measured with this amplifier is shown in Fig. 2. These experimental results were found in agreement with the theoretical relations (4) and (5).

By cooling the diode to liquid nitrogen temperature (78°K), the minimum amplifier noise temperature was reduced to 50° K, while the pump power necessary to maintain constant gain was decreased by 25 per cent. This was also found consistent with (4) and (5).

The authors are indebted to K. M. Johnson for the data presented in Fig. 2. and to M. R. Currie for helpful discussions.

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Rotating Loop Reflectometer for Waveguide*

The rotating inductive loop described by Tischer[1] has been successfully used by us as a reflectometer in S-band waveguide; limited access precluded the use of a slotted line or a directional coupler. However, considerable skill in adjusting the loop and compensating probe is called for when it is required to measure VSWR of the order of 1.01, and this setting is very easily disturbed. I therefore feel that a variant of the method requiring much less critical adjustment might be of some interest.

In principle, if a pure inductive loop is inserted through the wall of a waveguide and rotated, at a position where the magnetic field of the mode is circularly polarized, the signal induced in it will remain constant independent of its orientation. However, there is also an electric field at this position, and, in practice, a simple wire loop connected to a crystal detector behaves as a capacitive probe as well as an inductive loop; previous techniques have used a separate probe, critically coupled to the loop to cancel out the capacitive component of the signal.

Here the effect of pure inductance is achieved by closing the loop, through another loop which is then only inductively coupled to the mode in a second waveguide. One position satisfying this condition is at the center of a plane short circuit placed across a rectangular guide propagating the H_{01} or H_{011} modes. A capacitive probe inserted here perpendicular to the surface is completely decoupled from the electric field, but a loop can be orientated for maximum magnetic coupling. The device is illustrated in Fig. 1. Symmetry and careful positioning of the loops in both waveguides are necessary, and any higher modes excited in the secondary guide should be isolated from the detector.

* Received by the IRE, November 27, 1958.
[1] F. J. Tischer, "Rotatable inductive probe in waveguides," Proc. IRE, vol. 43, pp. 974–980; August, 1955.

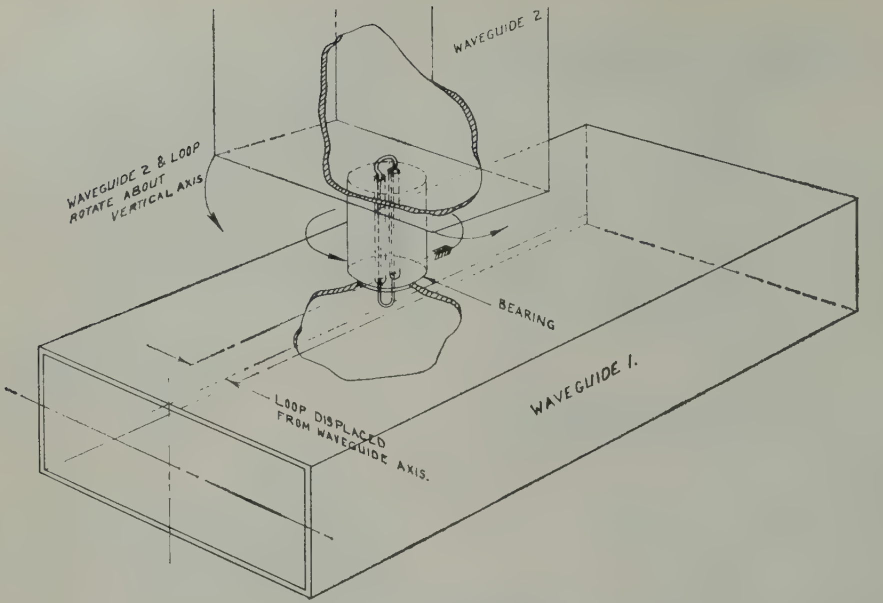

Fig. 1—Rotating loop reflectometer.

Nevertheless the adjustments are much less critical than for the earlier device. Thus, pure inductive coupling is obtained without need for a compensating probe, and, in addition, the loop may readily take various sizes and shapes according to the application.

The principle was confirmed with a makeshift model used as a reflectometer in rectangular guide. This gave, for example, an apparent VSWR of about 1.01 when a high precision slotted line instrument indicated a VSWR of 1.003. Unfortunately, very little development has been carried out, but with suitable refinements a greater accuracy should be obtained. A particularly valuable application is as a reflectometer for use in rectangular and circular waveguides of nonstandard sizes because only a small hole through the guide wall is required and the rotating mechanism may then be clamped in position.

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Gain Measurements on a Pulsed Ferromagnetic Microwave Amplifier*

The dependence of gain on pump power has been measured for a ferromagnetic microwave amplifier[1] using polycrystalline yttrium iron garnet. The amplifier operates in the "quasi-degenerate" mode in which one cavity serves as the resonant circuit for both the signal and the idle frequency fields. A second resonant cavity is used for the pump signal. Coupling between the signal-idle cavity and the external transmission line is obtained with a single capacitive probe, so that the device is a reflection-type negative-resistance amplifier. To reduce heating effects a pulsed source is used for the pump power; the device amplifies only when the pump is on.

With the CW signal f_s tuned to approximately one half the pump frequency, the power reflected from the signal-idle cavity was measured as a function of the peak pump power incident on the pump cavity. The reflected power was measured relative to the incident signal power by adjusting a precision attenuator to give the same amplitude of power at the input of a microwave receiver as that measured with a short inserted immediately in front of the cavity. The reflected power was allowed to build up for 1.5 μsec after the beginning of the pump pulse before it was measured. The power gain was defined as the ratio of the power reflected from the signal-idle cavity to the power incident on the cavity at the signal frequency. The pump power was measured with a directional coupler and a bolometer. The incident signal frequency power was varied from 2×10^{-7} watts to 2×10^{-3} watts for these measurements to test for possible signal saturation. At signal levels greater than 10^{-2} watts severe arcing occurred in the signal-idle cavity at high gains. The results of this experiment are shown in Fig. 1.

Using the expression of Suhl[1] for the negative quality factor (Q_s) of the ferrite one can derive the following expression for the gain of a reflection-type ferrite amplifier under transient conditions:

$$g(t) = 1 - \frac{4Q_L^2}{Q_eQ_u}\,\frac{\left(1-\dfrac{Q_u}{Q_L}\,\dfrac{P}{P_c}\right)}{\left(1-\dfrac{P}{P_c}\right)^2}\cdot\left[1-\frac{P}{P_c}e^{-Bt/2}\right]^2 \tag{1}$$

where Q_L, Q_e, and Q_u are, respectively, the loaded, external, and unloaded quality factors of the signal-idle cavity

$$\frac{1}{Q_L} = \frac{1}{Q_e} + \frac{1}{Q_u};$$

P is the pump power incident on the cavity; P_c is the value of pump power which gives infinite steady state gain (*i.e.*, the value of P which drives the precession angle to the critical value given by Suhl for electromagnetic operation[1]); t is the time elapsing after the beginning of the pump pulse; and B is the amplifier bandwidth, which for large gain is

$$B \doteq \frac{2\pi f_s}{Q_L}\,\frac{\left(1-\dfrac{P}{P_c}\right)}{\left(1+\dfrac{P}{P_c}\right)}. \tag{2}$$

In deriving (1) and (2) it has been assumed that the ferromagnetic precession angle is directly proportional to the amplitude of the field at the pump frequency; *i.e.*, high power saturation effects[2] have been neglected. If these effects were included, one would expect P/P_c to be replaced by some other more slowly varying function of P.

It was not possible to measure P_c by independent means in this experiment because of the pulsed operation of the device (see next paragraph) but a reasonable fit of (1) to the data can be obtained if one assumes $P_c = 5.0$ kw. The theoretical transient gain and steady-state gain are indicated in Fig. 1.

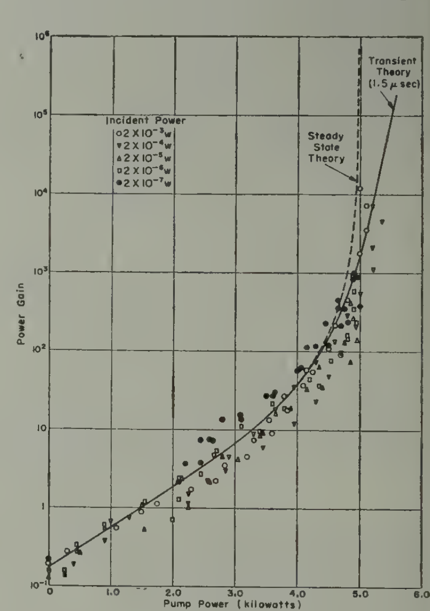

Fig. 1—Power gain as a function of pump power.

The rather wide scatter of the experimental points may have been caused by fluctuations in the pump frequency and power, and by phase instabilities resulting from the near equality of the signal and idle frequencies.

In an effort to determine the noise properties of the device the output power of the amplifier at one half the pump frequency was measured at various values of pump power with no signal input. On the assumption that the output power resulted from the amplification of noise present in the amplifier and in the external room temperature load, the

* Received by the IRE, December 11, 1958.
[1] H. Suhl, "Theory of the ferromagnetic microwave amplifier," *J. Appl. Phys.*, vol. 28, pp. 1225–1236; November, 1957.

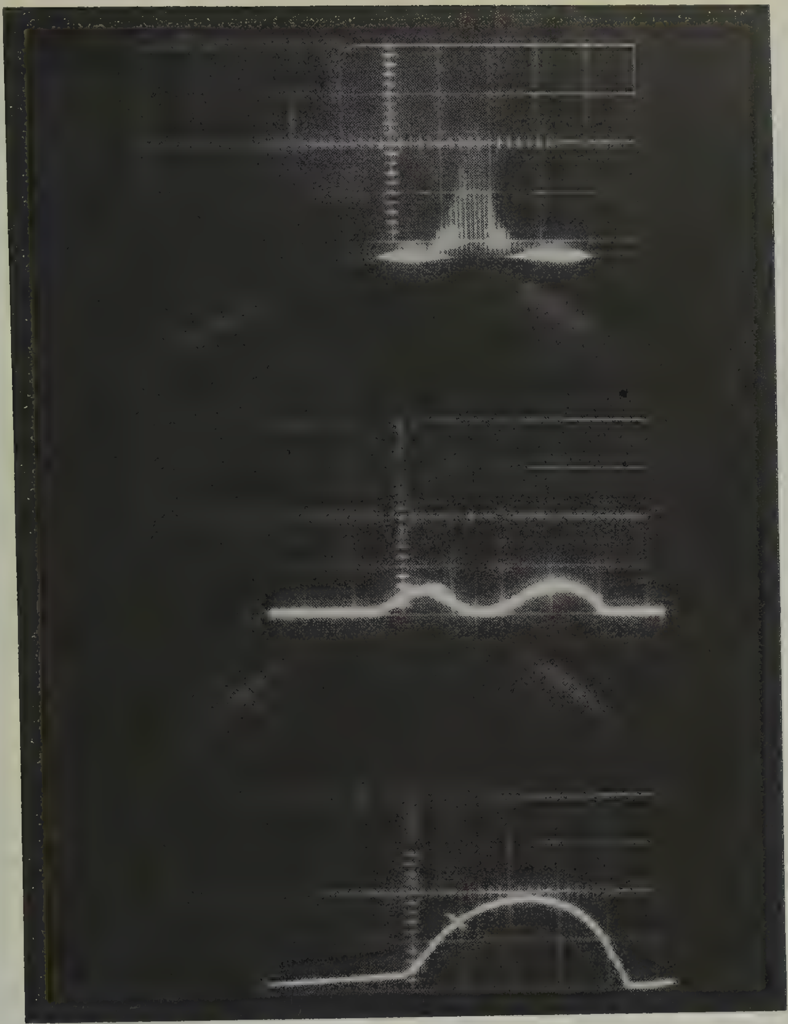

Fig. 2—Power reflected from ferromagnetic amplifier as a function of input signal frequency.

noise power of the device referred to the input was computed. The effective input noise power was found to be approximately 10^{-10} watts for values of pump power between 2 and 5 kw. It is believed that a major portion of this noise results from the pump pulse transients and that CW operation would reduce the effective input noise power to a much lower value.

The product of the bandwidth and the square root of the power gain was calculated from the measured values for power gains between 7 and 70 and had a nearly constant value of 15 mc.

In Fig. 2 the power reflected from the signal-idle cavity is shown under three different conditions. In all three of these pictures the horizontal scale is proportional to signal frequency and the vertical scale is proportional to signal power. The lower picture shows the signal power reflected from a short inserted just in front of the cavity; this is therefore the power incident on the cavity. The amplitude variation results from the variation in the signal klystron output power with frequency. In the center picture the short has been removed and the signal-idle cavity absorption is observed with the pump off. In the upper picture the peak pump power is 3.7 kw at constant frequency and the amplified signal from the ferrite amplifier is traced out as a function of signal frequency. One unit of the vertical scale of the upper picture corresponds to 10 units of the vertical scale for the other two pictures so that the power gain is about 20. The amplifier bandwidth for this gain is seen to be of the same order of magnitude as the bandpass of the signal-idle cavity.

Using a value of 50 gauss for the width of the ferromagnetic resonance, we compute $P_c \doteq 5 \times 10^2$ watts from the expression given by Suhl.[1] The discrepancy between the theoretical and the experimental values of P_c is believed to be caused in part by the saturation of the uniform precessional mode at the high values of pump power used in the device.[2]

The amplifier design is similar to that of Weiss,[3] and consists of a flattened coaxial resonant line in a rectangular cavity. The pump signal is fed into the rectangular cavity through a variable coupling hole. The signal-idle line is one-half wavelength long with two 0.125-inch diameter disks of polycrystalline yttrium iron garnet on opposite sides of the center conductor. These disks are each 0.050 inch thick. At low X-band power levels the resonance width for spheres of this material is 50 gauss, but at the high pump power levels used in the amplifier, saturation effects[2] increased the width of the resonance so that operation was obtained over a range of more than 200 gauss.

The experimentally determined quality factors of the signal-idle cavity were $Q_L = 3 \times 10^2$ and $Q_u/Q_L = 3.39$. The pump frequency was 9.26 kmc with a pump pulse duration of 2.1 μsec and a pulse repetition rate of 350 cycles. The signal frequency was approximately 4.63 kmc. The pump cavity was very nearly matched to the transmission line and had a loaded Q of approximately 10.

The authors wish to thank P. Bailey of the Westinghouse Magnetic Materials Development Laboratory for preparing the samples of yttrium iron garnet.

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[2] H. Suhl, "The nonlinear behavior of ferrites at high microwave signal levels," Proc. IRE, vol. 44, pp. 1270–1284; October, 1956.

[3] M. T. Weiss, "Solid state microwave amplifier and oscillator using ferrites," *Phys. Rev.*, vol. 107, p. 317; July 1, 1957.

Parametric Devices Tested for Phase-Distortionless Limiting*

Two parametric devices using variable capacitance diodes have been successfully tested for phase-distortionless limiting. One of the devices is shown schematically in Fig. 1. An input signal at 100 mc is combined

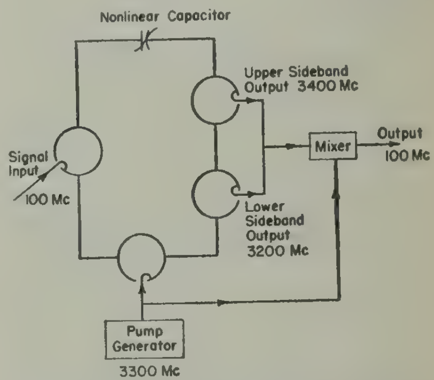

Fig. 1—Diagram of the 100-mc parametric amplifier. The circles represent resonant cavities which are coupled together in series with a nonlinear capacitor.

with a pump input at 3300 mc to produce power at both the upper and the lower sidebands. This power is then mixed in conventional fashion with power from the 3300-mc pump to give an output at 100 mc, the frequency of the input. It is this output which exhibits good limiting characteristics. As shown in Fig. 2, the output remains constant to within $\pm\frac{1}{2}$ db over a range of input

* Received by the IRE, December 15, 1958. This work was carried on in cooperation with the Air Force Cambridge Research Center under Contract AF19-(604)1847.

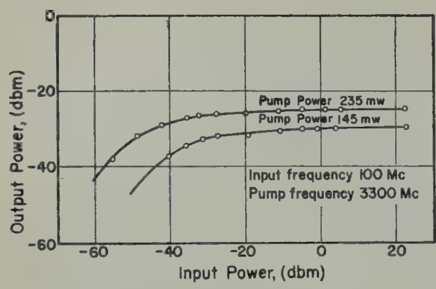

Fig. 2—Saturation characteristics of the 100-mc amplifier.

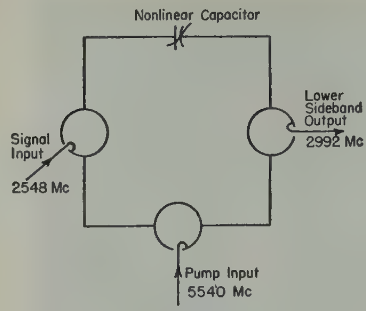

Fig. 3—Diagram of the S-band parametric frequency converter.

of greater than 45 db. Some of the limiting may be due to components in the system other than those which produce the parametric effects. The amount of phase distortion (*i.e.*, change in the phase of the output as a function of input power) has been difficult to measure for this device. Preliminary measurements have indicated little or no phase distortion.

The other parametric device is an S-band frequency converter with gain in which the output is at the lower sideband frequency. The schematical diagram and the measured limiting characteristics are presented in Figs. 3 and 4, respectively. The output remains constant to within $\pm 1\frac{1}{2}$ db over a range of input of more than 20 db. Here the limiting is solely a consequence of the parametric action. The phase characteristics have been measured to much greater accuracy for this device than for the preceding one. A plot of output phase vs input power gives a curve of nearly constant slope, representing a change of output phase of about one degree per decibel of change in input power.

In searching for an explanation of the limiting behavior exhibited in the two devices it became apparent from the theory that a limiting mechanism exists in connection with simple parametric frequency conversion. This mechanism is at least partially responsible for the limiting obtained from the first device and plays a major part in the limiting obtained from the second. The mechanism can be explained from a first-order analysis of the simplified equivalent circuit shown in Fig. 5. For simplicity the resonant circuits are assumed to be lossless. As shown in the figure each one is loaded by either a shunt conductance from an input generator or the conductance of the output load.

It is assumed that the loaded Qs are sufficiently high for each resonant tank to present a short circuit to voltages at the resonant frequencies of the other two. The coupling capacitor is a nonlinear element whose capacitance changes with voltage as follows:

$$C = C_0 + DV.$$

The steady capacitance C_0 can be regarded as a part of each of the tanks as shown in Fig. 5. The subscripts 1, 2, and p refer to input tank quantities, idling or output tank quantities, and pumping tank quantities, respectively.

When resonance operation is assumed, the conversion gain from an input signal at

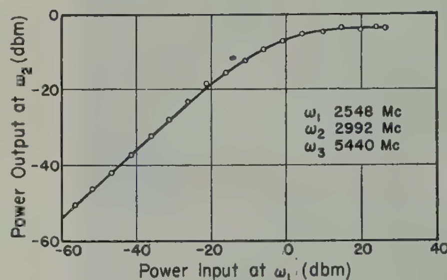

Fig. 4—Saturation characteristics of the S-band parametric frequency converter.

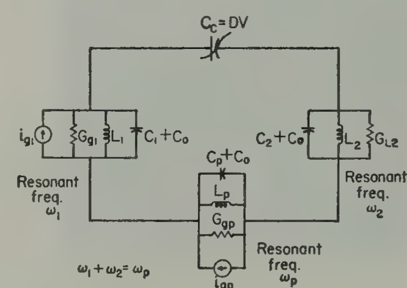

Fig. 5—Equivalent lumped circuit for the analysis of the parametric frequency converter.

ω_1 to an output signal at ω_2 is[1]

$$\text{gain} = \frac{4 \dfrac{\omega_2}{\omega_1} G_{g1} G}{(G_{g1} - G)^2} \tag{1}$$

where

$$G = \frac{\omega_1 \omega_2 D^2 V_p^2}{G_{L2}}. \tag{2}$$

The limiting mechanism becomes apparent through an examination of the equations relating the voltage across the pumping resonant circuit, V_p, to the voltage across the input resonant circuit, V_1. The following two equations present this relationship[1]

$$V_p = \frac{i_{gp}}{G_{gp} + \dfrac{\omega_p \omega_2 D^2 V_1^2}{G_{L2}}} \tag{3}$$

$$V_1 = \frac{i_{g1}}{G_{g1} - \dfrac{\omega_1 \omega_2 D^2 V_p^2}{G_{L2}}}. \tag{4}$$

In the operation of this device as a limiter i_{gp} is held constant as i_{g1} is varied. Eqs. (3) and (4) show that feedback between V_p and V_1 exists which tends to hold V_1 constant even though i_{g1} is increased. For example let us assume that increasing i_{g1} produces an increase in V_1 as (4) suggests. If V_1 increases, then according to (3), V_p should decrease. But a smaller V_p in (4) tends to reduce V_1 which is contrary to the assumed effect of increasing i_{g1}. The stabilizing mechanism becomes more and more effective as

$$\frac{\omega_1 \omega_2 D^2 V_p^2}{G_{L2}}$$

approaches closer and closer to G_{g1} in magnitude. This stabilization is reflected in a stabilization of the voltage V_2 across the output tank. In this fashion a limiting mechanism exists which tends to keep the output power $V_2^2 G_{L2}$ at a constant value even though the available input power $i_{g1}^2/4G_{g1}$ is caused to vary.

By using a general power relationship from Manley and Rowe[2] we can express the output power conveniently as a function of both V_1 and V_p. Hence

$$V_2^2 G_{L2} = P_{\text{out}} = \frac{\omega_2}{\omega_p} P_{\text{pump}}$$

$$= \frac{\omega_2}{\omega_p} V_p^2 \frac{\omega_2 \omega_p D^2 V_1^2}{G_{L2}}. \tag{5}$$

Eqs. (3)–(5) permit the calculation of output power as a function of available input power. This was done using as the values for the various parameters those corresponding to the experimental conditions for the second device. The resulting theoretical curve did exhibit the limiting characteristic. However the range of limiting for the theoretical curve was not as great as that obtained experimentally. The theory is now being extended to include higher order terms and also to consider the effects of possible variations in the biasing of the diode. (In all of the experiments the diodes were self-biased.)

The phase characteristics are also being investigated theoretically. It is presently believed that the higher order terms and the biasing effects represent sources of phase distortion. The experimental data indicate that their importance cannot be too great.

The authors wish to acknowledge helpful discussions with D. A. Linden, who suggested the physical nature of the limiting mechanism, H. Heffner, who contributed to the mathematical analysis, and K. L. Kotzebue, who did an initial large signal analysis.[3]

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[1] H. Heffner and G. Wade, "Gain, bandwidth, and noise characteristics of the variable-parameter amplifier," *J. Appl. Phys.*, vol. 29, pp. 1321–1331; September, 1958.

[2] J. M. Manley and H. E. Rowe, "Some general properties of nonlinear elements—Part I. General energy relations," Proc. IRE, vol. 44, pp. 904–913; July, 1956.

[3] K. L. Kotzebue, "A Semiconductor-Diode Parametric Amplifier at Microwave Frequencies," Electronics Labs., Stanford University, Stanford, Calif., Tech. Rep. No. 49; November 4, 1958.

DC Characteristics of a Junction Diode*

Moll[1] has pointed out the problems involved in comparing theory and experiment for p-n junction diodes. Thus, the well-known expression of Shockley, $I=I_0(e^{\alpha^* V}-1)$ with $\alpha^*=q/kT$, holds only for heavy doping or very small applied voltages. For an intrinsic base, Spenke[2] shows that the largest value for α^* is $q/2kT$. For intermediate doping, α^* is certainly smaller than q/kT, and for large forward bias one again has $q/2\,kT$, as quoted by Moll. We wish, therefore, to describe a diode for which, by including the field, all these relationships can be connected analytically.

This is a planar, narrow-base, abrupt-junction diode with unequal doping (injection into base only). For later use two restrictions are restated:

width of base, $W \ll (D_p \tau_p)^{1/2}$ (neglect of recombination) and (1)

number of ionized acceptors in emitter region, $N_a \gg$ number of ionized donors in base, N_d. (2)

Fig. 1 is a schematic diagram of the diode. We use the boundary conditions

$$p = p_0 e^{\alpha V_1}, \quad x = 0; \quad \alpha = q/kT \qquad (3)$$

$$p = p_0, \qquad x = W. \qquad (4)$$

Because of (1) and (2), the net electron current in the base can be set equal to zero, and the hole current, J_p, must be constant. Using the definitions of electron and hole current

$$J_n = q\mu_n nE + qD_n \nabla n = 0 \qquad (5)$$

$$J_p = q\mu_p pE - qD_p \nabla p \qquad (6)$$

together with the neutrality condition

$$p - n + N_d = 0 \qquad (7)$$

we have for the field in the base,

$$E = \frac{J_p + qD_p \nabla_p}{q\mu_p p} \qquad (8)$$

and also

$$E = -\frac{D_p p}{\mu_p(p + N_d)} \cdot \qquad (9)$$

Solving for ∇_p between these two equations:

$$\nabla_p = -\frac{Jp}{qDp}\left(\frac{p+N_d}{2p+N_d}\right). \qquad (10)$$

Integrating (10) and using (3) and (4), the total current obtained is

$$J = J_p = qD_p N_d/W \left[2p_0/N_d(e^{\alpha V_1}-1) - \ln\left(\frac{N_d + p_0 e^{\alpha V_1}}{N_d + p_0}\right)\right]. \qquad (11)$$

These relations have been derived by Rittner[3] in connection with transistor theory.

* Received by the IRE, January 9, 1959.

[1] J. L. Moll, "The evolution of the theory for the voltage-current characteristic of p-n junctions," Proc. IRE, vol. 46, pp. 1076–1082; June, 1958.

[2] E. Spenke, "Forward and back characteristics of a P-I-metal rectifier," *Z. Naturf.*, vol. 11a, no. 6, pp. 440–456; 1956.

[3] E. S. Rittner, "Extension of the theory of the junction transistor," *Phys. Rev.*, vol. 94, pp. 1161–1171; June 1, 1954.

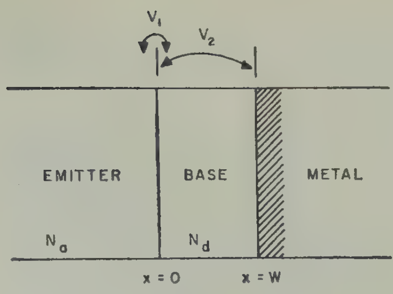

Fig. 1—The narrow-base diode.

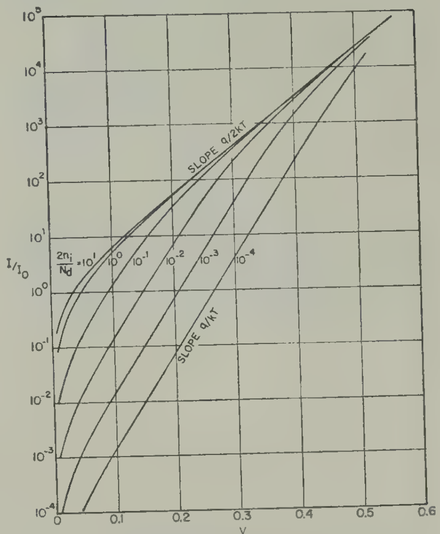

Fig. 2—Forward characteristics of the narrow-base diode.

Eliminating ∇p between (8) and (9), we have

$$E = \frac{J_p}{q\mu_p(2p + N_d)} \qquad (12)$$

for the field in the base region. The voltage drop across the base, V_2 is then given by

$$V_2 = \int_0^W E(x)dx = \int_{P_0 e^{\alpha V_1}}^{P_0} E(p)\left(\frac{dx}{dp}\right)dp \qquad (13)$$

which can be evaluated to give

$$V_2 = 1/\alpha \ln\left(\frac{N_d + p_0 e^{\alpha V_1}}{N_d + p_0}\right). \qquad (14)$$

Since $V_{\text{total}} \equiv V = V_1 + V_2$, V_2 can be eliminated in (14) and solve for V_1 to get

$$\alpha V_1 = -N_d/2p_0\{1 - [1 + 4p_0/N_d(p_0/N_d + 1)e^{\alpha V}]^{1/2}\}. \qquad (15)$$

The equilibrium concentration of holes in the base, p_0, can be written as

$$p_0 = -N_d/2\{1 - [1 - (2n_i/N_d)^2]^{1/2}\} \qquad (16)$$

where (7) and the equilibrium relation $np = (n_i)^2$ are used. Substituting (15) and (16) into (11) finally leads to

$$I = I_0/a\left\{(1 + a^2 e^{\alpha V})^{1/2} - (1 + a^2)^{1/2} - \ln\left[\frac{1 + (1 + a^2 e^{\alpha V})^{1/2}}{1 + (1 + a^2)^{1/2}}\right]\right\} \qquad (17)$$

where $I_0 = 2qD_p n_i A/W$, $A = 2\,n_i/N_d$, A is the junction area, and n_i is the intrinsic concentration of electrons in the material.

It is readily seen that (17) reduces to $I = I_0 a/4(e^{\alpha V}-1)$ for small values of the parameter a. This is also the result obtained by direct application of the Shockley theory to a similar diode with heavy base doping.

For large values of a, (17) reduces to $I = I_0(e^{\alpha V/2}-1)$, the result of Spenke for intrinsic base material.

The gradual change in α^* from q/kT to $q/2kT$ is illustrated in Fig. 2, which shows (17) for a series of values of the parameter a. For large V, α^* is seen to approach $q/2kT$ for any doping.

The range of validity of this equation can be estimated from the condition

$$e^{\alpha V/2} \ll N_a/n_i.$$

Valuable discussions with A. Brodzinsky are gratefully acknowledged.

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Suppression of Undesired Radiation —Directional HF Antennas and Associated Feed Lines*

Concerning RF transmission lines of the open type,[1] the statement "Yet, the provisions presently made for detecting such deficiencies are totally inadequate in almost all installations, both commercial and military, to the knowledge of the author," deserves some comment. The "deficiencies" referred to are unbalanced components on the line.

In our point-to-point HF transmitting installations, it has been the practice to use a balance-indicating voltmeter, similar to the type shown in Fig. 8 by Brueckmann, for the past nine years. Our circuit has the additional feature of being tunable over the 4–26 mc range to minimize pick-up from lines and frequencies other than that being measured. It is known as a type BNY-1197. The original design was by G. T. Royden.

In its portable form, this balance indicator consists of a small metal box with a wooden handle and two hooks for attachment to any point on the line. The need for such a device was recognized long ago, and all of our major installations are equipped with one or more of these units.

In discussing the demerits of using RF ammeters, the author also states, "Yet, two ammeters are presently standard equipment for monitoring." Here again, we wish to differ.

In the latest Mackay 10-kw and 30-kw transmitters, the output monitor consists of a built-in balance indicator and a built-in relative output voltage indicator. There are

* Received by the IRE, August 26, 1958.

[1] H. Brueckmann, "Suppression of undesired radiation of directional HF antennas and associated feed lines," Proc. IRE, vol. 46, pp. 1510–1516; August, 1958.

no ammeters. Transmitters so equipped have been in operation for several years.

The transmitter balance indicator plus the portable unit are essential in keeping our lines reasonably well balanced. Yet I would be the first to admit that, even so, open lines leave a great deal to be desired. In a large installation there are so many lines and so many bends and irregularities before the antenna feed point is reached, that the maintenance of truly balanced conditions throughout the system becomes difficult.

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A UHF Ruby Maser*

A ruby maser has been operated at signal frequencies tunable over the range 380–450 mc. In the experimental arrangement, the magnetic levels $M=+\frac{1}{2}$ and $M=-\frac{1}{2}$ of paramagnetic pink ruby were used for the signal, and X-band pumping was carried out between levels $M=+\frac{3}{2}$ and $M=+\frac{1}{2}$. At the required low magnetic field strength of approximately 70 oe, maser operation was observed to be only slightly dependent on the angle between the dc magnetic field and the crystal axis; amplification and oscillation were obtained for any angle of orientation, with optimum performance at 90°. The observed slight dependence of the maser action on the angle of orientation at the low fields used here is expected on theoretical grounds.

In the mode of operation used, the X-band pumping frequency (approximately 11.8 kmc) is not a critical parameter so that tuning of the signal frequency over the range 380–450 mc required neither adjustment of the pump source nor variation of the magnetic field with regard to magnitude or direction. This feature makes the UHF ruby maser especially suitable for applications in which tuning is desired.

The ruby crystal (4 cm^3 in volume) was located at the center of a teflon loaded cavity which was excited in a TE_{011} pump mode. The construction and tuning of the signal frequency lumped circuit was carried out in a manner similar to that used by Kingston[1] in the design of a potassium-chromicyanide maser. The dc magnetic field was supplied by a small permanent magnet in which the field strength was adjusted by varying the gap between the pole faces.

As an amplifier, typical operational data included a gain of 15 db and a bandwidth of 100 kc with the maser at a temperature of 1.7°K. As an oscillator, the power output was less than 1 μw.

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* Received by the IRE, December 15, 1958.
[1] R. H. Kingston, "A UHF solid-state maser," PROC. IRE, vol. 46, p. 916; May, 1958.

Incoherent Scattering of Radio Waves by Free Electrons with Applications to Space Exploration by Radar*

Gordon[1] has discussed the use of a powerful radar to measure electron temperatures of the ionosphere. An interesting adjunct to these experiments would be to investigate Bailey's suggestion[2] of controlling the electron temperature of the ionosphere by means of radio waves of frequency 1.2 mc, the gyro frequency of ionospheric electrons. A section of the ionosphere could be illuminated by 1.2 mc radiation and the increased width in the frequency spectrum of the back scattered radar signal measured.

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* Received by the IRE, December 1, 1958.
[1] W. E. Gordon, "Incoherent scattering of radio waves by free electrons with applications to space exploration by radar," PROC. IRE, vol. 46, pp. 1824–1829; November, 1958.
[2] V. A. Bailey, "Control of the ionosphere by means of radio waves," *J. Atmos. Terres. Phys.*, vol. 12, No. 13, pp. 216–217; 1958.

WWV Standard Frequency Transmissions*

Since October 9, 1957, the National Bureau of Standards radio stations WWV and WWVH have been maintained as constant as possible with respect to atomic frequency standards maintained and operated by the Boulder Laboratories, National Bureau of Standards. On October 9, 1957, the USA Frequency Standard was 1.4 parts in 10^9 high with respect to the frequency derived from the UT 2 second (provisional value) as determined by the U. S. Naval Observatory. The atomic frequency standards remain constant and are known to be constant to 1 part in 10^9 or better. The broadcast frequency can be further corrected with respect to the USA Frequency Standard, as indicated in the table below. This correction is *not* with respect to the current value of frequency based on UT 2. A minus sign indicates that the broadcast frequency was low.

WWV FREQUENCY*

January, 1959 1600 UT	Thirty-Day Moving Average Seconds Pulses on 15 MC, Parts in 10^{10}†
1	−30
2	−30
3	−30
4	−30
5	−30
6	−30
7	−30
8	−29
9	−29
10	−29
11	−29
12	−29
13	−29
14	−28
15	−28
16	−28
17	−28
18	−28
19	−28
20	−28
21	−28
22	−28
23	−28
24	−28
25	−28
26	−28
27	−28
28	−28
29	−29
30	−29
31	−29

* Received by the IRE, February 24, 1959.
* WWVH frequency is synchronized with that of WWV.
† No adjustment was made in the control oscillator at WWV this month.

The WWV and WWVH time signals are synchronized; however, they may gradually depart from UT 2 (mean solar time corrected for polar variation and annual fluctuation in the rotation of the earth). Corrections are determined and published by the U. S. Naval Observatory.

WWV and WWVH time signals are maintained in close agreement with UT 2 by making step adjustments in time of precisely plus or minus twenty milliseconds on Wednesdays at 1900 UT when necessary; one retarding time adjustment was made during this month at WWV and WWVH on January 28, 1959.

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History of the Problem of Conversion of Heat to Electricity by Thermionic Emission*

Hatsopoulos and Kaye state in their paper[1] that the first electronic and thermodynamic study on the conversion of heat to electricity by thermionic emission was given by Hatsopoulos.[2]

The first properly documented scientific study of the fundamental problems of the conversion of heat to electricity by thermionic emission is that of Schlichter,[3] which was finished sometime prior to August 2, 1914.

Schlichter states[4] (literal translation): "We arrive at the following result: The total practical efficiency of the simple thermionic element is very small, if one uses Platinum and similar materials as electrodes on account of their low electron emission with respect to the radiation loss. It is not possible to reduce this radiation loss at a given emissivity in proportion to this emissivity. The practical efficiency could be increased only by selection of an electrode material with a sufficiently high melting

* Received by the IRE, January 5, 1959.
[1] G. N. Hatsopoulos and J. Kaye, "Analysis and experimental results of a diode configuration of a novel thermolectron engine," PROC. IRE, vol. 46, pp. 1574–1579; September, 1958.
[2] G. N. Hatsopoulos, "The thermoelectric engine," Sc.D. dissertation, M.I.T., Cambridge, Mass.; May, 1956.
[3] W. Schlichter, "Spontaneous emission from glowing metals and the thermionic element," (in German), inaugural dissertation, University of Göttingen, Ger., August, 1915. Published by Barth, Leipzig, Ger.; 1915. See also *Ann. Phys.*, vol. 47, pp. 573–640; 1915.
[4] *Ibid.*, inaugural dissertation, pp. 97–98.

point and extraordinary higher electron emission than that of Platinum. There does not exist a material of this kind among those which have been so far investigated."

"It does not seem to be entirely impossible to prepare such a material artificially by treatment of its surface. If one should succeed in this manner to reduce the constant ϕ to about one third of its value for Platinum maintaining the same constant A, a material with the required high electron emission would be produced. Thus the technical problem would also be solved. Neglecting the radiation loss, economically working thermionic elements are definitely possible within the limits set by the remaining physical laws."

It is evident from the text of Schlichter's dissertation (1a and 1b) that he was not familiar with the work of Child (1911),[5] Langmuir (1913),[6] and Schottky (1914).[7]

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Author's Comment[8]

Professor Hatsopoulos and I seriously doubt that we should reply to the statements raised by Dr. Miller. Let me indicate that we never claimed that we were the first ones to think of thermionic emission as a new means of conversion of heat directly into electricity. We clearly stated[9] that the *idea was quite old but that the first detailed electronic and thermodynamic analysis on this subject was given in a quantitative fashion by one of the authors in 1956.* Miller's remarks indicate that his first reference, Wilhelm Schlichter, discussed the idea only in a qualitative fashion without giving any detailed electronic and thermodynamic analysis.

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[5] C. D. Child, "Discharge from hot CaO," *Phys. Rev.*, vol. 32, pp. 492–511; May, 1911.

[6] I. Langmuir, "The effect of space charge and residual gases on thermionic currents in high vacuum," *Phys. Rev.*, vol. 2, pp. 402–403; 450–486; December, 1913.

[7] W. Schottky, "Action of space charge in thermionic currents in high vacuum," (in German), *Phys. Zeit.*, 1914, vol. 15, pp. 526–528; 1914.

[8] Received by the IRE, January 21, 1959.

[9] *Op. cit.*, p. 1575.

An Approximation of Transient Response from Frequency Response Data*

The determination of a system transient response, if the frequency response is known, has been studied in many publications.

A method that simplifies Dawson's[1] procedure is presented here.

If the system frequency response is known as

$$H(j\omega) = A(\omega) + jB(\omega), \qquad (1)$$

then the impulse response is

$$h(t) = \frac{-2}{\pi} \int_0^{\infty_0} B(\omega) \sin \omega t d\omega. \qquad (2)$$

Usually $B(\omega)$ is quite small for large values of ω such that the approximation,

$$B(\omega) \cong 0 \qquad \omega > \omega_0,$$

will not change the integral for $h(t)$.

Therefore,

$$h(t) \cong \frac{-2}{\pi} \int_0^{\omega_0} B(\omega) \sin \omega t d\omega. \qquad (3)$$

Also $B(\omega)$ can be described by a Fourier series for $0 < \omega < \omega_0$ as

$$B(\omega) = \sum_{n=1}^{\infty} \sin \frac{n\omega}{\omega_0} \qquad (4)$$

where

$$b_n = \frac{2}{\omega_0} \int_0^{\omega_0} B(\omega) \sin \frac{n\pi\omega}{\omega_0} d\omega. \qquad (5)$$

But, using the approximate equation for $h(t)$,

$$b_n = -\frac{\pi}{\omega_0} h\left(\frac{n\pi}{\omega_0}\right) \qquad (6)$$

or

$$h\left(\frac{n\pi}{\omega_0}\right) = -\frac{\omega_0}{\pi} b_n$$

$$h\left(\frac{nT_0}{2}\right) = \frac{-2b_n}{T_0}. \qquad (7)$$

Thus the determination of the transient response at discrete values of time, $t = nT_0/2$, reduces to the determination of Fourier series coefficients, which is relatively simple and well tabulated.

The example worked out by Dawson was a closed looped unity feedback system with a forward transfer function,

$$G(j\omega) = 5(1 + j\omega)^2/((j\omega)^3(1 + j0.05\omega)^2).$$

Dawson determined the coefficients b_n to be (assuming $\omega_0 = 17.6$),

$$b_1 = -0.781 \qquad b_2 = -0.519$$

$$b_3 = -0.66 \text{ and } b_4 = 0.134.$$

Setting these values in (7) yields,

t	$h(t)$	$h_a(t)$
.179	4.38	4.39
.358	2.91	2.87
.536	.37	.30
.715	−.75	−.77

The values $h_a(t)$ are determined by the inversion of the closed loop transfer function.

$$h_a(t) = 0.09e^{-.726t} - 6.36e^{-2.8t} + 3.58e^{-28.13t} + 22.76e^{-4.17t} \sin(4.18t + 6.8°).$$

Comparison of the two columns shows the approximation to be quite close to actual values.

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* Received by the IRE, January 16, 1959.

[1] C. H. Dawson, "Approximation of transient response from frequency response data," *Trans. AIEE*, vol. 72, pt. II, pp. 289–291; 1953.

Space-Charge Capacitance of a *P-N* Junction*

When estimating the space-charge capacitance of a *p-n* junction, the space-charge density is often assumed to follow a definite function of position (that of excess donor density) for an appropriate distance. It always turns out that the incremental capacitance per unit cross section is ϵ/d, where ϵ is the dielectric constant and d the breadth of the space-charge layer, just like the capacitance of the corresponding parallel-plate condenser.

Though often implied, this is no coincidence. Given any fixed curve c for the space-charge density (see Fig. 1) the incremental charges $\pm dq$ will add at the boundaries of the space-charge region and thereby cause an additional field and hence an additional potential difference equal to that produced by these charges on condenser plates separated by the distance d.

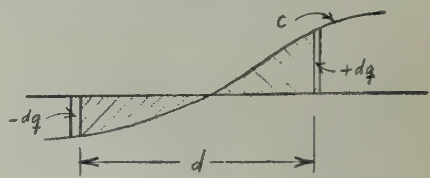

Fig. 1

HELGE JØRSBOE
The Technical University of Denmark
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* Received by the IRE, December 18, 1958.

On Direct Coupled Transistor Amplifier Stages*

A transistor is used frequently to perform the function of current amplification. If a transistor current amplifier must have a relatively constant current amplification to zero frequency, reactive elements cannot be used in the coupling or bias circuits. In this case high current amplification and good stability of the quiescent operating point with respect to temperature variations become inconsistent requirements unless temperature-sensitive elements are used to enhance the thermal stability of the circuit.

Temperature instability of the operating point in a transistor amplifier stage has two basic sources. First, the conductance of the emitter junction increases with increasing temperature.[1] Second, the collector cut-off current increases exponentially with increasing temperature. These effects will result in a temperature-sensitive operating point unless the emitter current is made substantially independent of the parameters of the transistor. In practice this is usually accom-

* Received by the IRE, August 7, 1958; revised manuscript received, November 20, 1958.

[1] J. S. Schaffner and R. F. Shea, "The variation of the forward characteristics of junction diodes with temperature," PROC. IRE, vol. 43, p. 101; January, 1955.

plished by making the resistance of the emitter-to-collector path in the external circuit large compared to the resistance of the base-to-collector path. This stabilization technique reduces the current amplification of a transistor amplifier stage when the transistor is used in the common-emitter or common-collector connection.

The purpose of this note is to derive a result of general significance regarding the relationship between the current amplification and the thermal stability factor of a transistor dc amplifier which is biased with linear temperature-invariant resistors. It is assumed that the external resistance in series with the emitter is large enough to overcome the effects of thermally-induced variations in the emitter conductance. Therefore, temperature instability is caused only by variations in the collector cut-off current, I_{CO}.

Consider a transistor which is connected *arbitrarily* in a *resistive* network as shown in Fig. 1. Note that no assumptions are made regarding the detailed connections of the transistor in the circuit. The results of the following analysis will thus apply to any transistor amplifier stage which does not contain reactive elements or temperature-sensitive resistors. The analysis is made in terms of incremental variations of the operating point. Thus the circuit can be regarded as possessing no sources except for the incremental input source and the dependent generator of the transistor model. With regard to the transistor, assume that the incremental collector current is given by

$$i_c = \alpha i_e + i_{co} \tag{1}$$

where α is the short-circuit incremental current transfer ratio of the transistor, i_e is the incremental emitter current, and i_{co} is the increment in collector cut-off current caused by a temperature variation. Note that this assumes that the reverse collector conductance is negligibly small. Also, as discussed above, assume that the emitter current is independent of temperature-induced variations in the emitter conductance because of the presence of external series resistance.

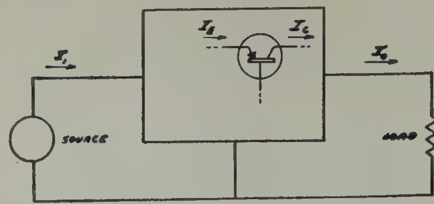

Fig. 1.

If i_1 and i_c are regarded as independent variables and i_e and i_o as dependent variables, then

$$i_o = A_{o1} i_1 + A_{oc} i_c, \tag{2}$$

$$i_e = A_{e1} i_1 + A_{ec} i_c \tag{3}$$

where

$$A_{o1} = \frac{i_o}{i_1}\Big|_{i_c=0} \qquad A_{oc} = \frac{i_o}{i_c}\Big|_{i_1=0}$$

$$A_{e1} = \frac{i_e}{i_1}\Big|_{i_c=0} \qquad A_{ec} = \frac{i_e}{i_c}\Big|_{i_1=0}$$

are current transfer ratios. These ratios can be computed by removing the transistor from the circuit, replacing the emitter-to-base junction by a short circuit, and the collector-to-base junction by a current generator. The four ratios thus defined are functions of the resistances of the circuit and therefore have magnitudes which are at most unity.

If (1) is substituted into (2) and (3), and if the resulting equations are solved for i_o, the result is

$$i_o = i_1\left[\frac{A_{oc}}{1 - A_{ec}\alpha}\alpha A_{e1} + A_{o1}\right] + i_{co}\left[\frac{A_{oc}}{1 - A_{ec}\alpha}\right]. \tag{4}$$

The first factor in brackets is the current amplification of the amplifier

$$A_i = \frac{i_o}{i_1}\Big|_{i_{CO}=0} = \frac{A_{oc}}{1 - A_{ec}\alpha}\alpha A_{e1} + A_{o1}, \tag{5}$$

and the second factor in brackets is the thermal stability factor as conventionally defined:[2]

$$S = \frac{i_o}{i_{co}}\Big|_{i_1=0} = \frac{A_{oc}}{1 - A_{ec}\alpha}. \tag{6}$$

Thus, it follows that

$$A_i = S\alpha A_{e1} + A_{o1}. \tag{7}$$

Since A_{o1} and A_{e1} have magnitudes which are at most unity, and since for a junction transistor α is less than unity, the incremental current amplification cannot exceed the thermal stability factor by more than one:

$$|A_i| \leq |S| + 1. \tag{8}$$

The thermal stability factor S is a measure of the drift in output current per unit drift in I_{CO} and thus must be made small for direct-coupled amplifiers when a drift in the operating point of one stage is a portion of the input to the next stage. This analysis shows that the current amplification of a transistor amplifier stage is severely restricted in magnitude if a small stability factor is required.

It is important to note that this analysis is for a single-transistor amplifier stage and does not apply if temperature-sensitive elements such as thermistors, semiconductor diodes, or other transistors are used to compensate for variations in I_{CO}. Also it does not apply directly in the case of cascaded stages when drift of one stage can be used to cancel a portion of the drift of the next stage.

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[2] R. F. Shea, "Principles of Transistor Circuits," John Wiley and Sons, Inc., New York, N. Y., p. 99; 1953.

Contributors

L. Ensing (M'58) was born in Schoonebeek, The Netherlands, on October 30, 1919. He received the diploma in electrical engineering from the Technical College of Rotterdam in 1941.

L. Ensing

In 1941 he joined the Philips Research Laboratories, Eindhoven, where he was occupied with the development of vacuum valves from 1941 to 1943, and with the development of electronic measuring instruments from 1943 to 1953. Since 1953 he has been employed at the Royal Dutch Shell Laboratory, Delft, as a research engineer on the development of electronic instruments for company use.

Mr. Ensing is a Netherlands Professional Engineer and a member of the NRG (Netherlands Association for Radio Engineering), the K.I.v.I. (Royal Institute of Engineers), and the International Association for Analog Computation.

❖

W. J. Grubbs received the B.S. degree in electrical engineering from the University of Kentucky, Lexington, Ky., in 1951, and joined the Bell Telephone Laboratories, Murray Hill, N. J., that same year. He completed the Bell Laboratories Communications Development Training Program in 1954, and is now working towards the M.S. degree in physics at Stevens Institute of Technology, Hoboken, N. J.

W. J. Grubbs

For some time after joining Bell Laboratories, he worked on the design of ferrite core inductors. Since 1956, he has been engaged in the design and fundamental development of solid-state devices.

J. M. L. Janssen (SM'56) was born in nhem, The Netherlands, on September 14, 18. He attended the Technological University of Delft, where he received the physical engineering degree in 1941. From 1942 to 1949 he worked with Philips Research Laboratories, Eindhoven, on the development of electronic measuring equipment.

J. M. L. Janssen

In 1949 Mr. Janssen joined the Royal Dutch Shell Laboraory, Delft, where he was placed in charge f the Measurement and Control Departnent the following year.

In 1957 he was transferred to the Bataafche Petroleum My., The Hague, as head of heir Data Processing and Computer Center.

❖

J. E. Rowe (A'51–M'55) was born in Highland Park, Mich., in 1927. He received he B.S. degree in electrical engineering and in mathematics in 1951, the M.S. degree in electrical engineering in 1952, and the Ph.D. degree in electrical engineering in 1955, all from the University of Michigan, Ann Arbor.

J. E. Rowe

Since 1951 he has been associated with the University of Michigan Research Institute, engaging in fundamental research on microwave systems, microwave devices, and electromagnetic field theory. Formerly a lecturer and assistant professor of electrical engineering, he is now an associate professor of electrical engineering and head of the Electron Physics Laboratory at the University of Michigan.

Dr. Rowe is a member of the American Institute of Electrical Engineers, the American Mathematical Society, and the Society for Industrial and Applied Mathematics. He is also a member of Sigma Xi, Phi Kappa Phi, Tau Beta Pi, and Eta Kappa Nu.

Bruno Schneider was born on August 6, 1929, in Zürich, Switzerland. He studied electrical engineering at the Swiss Federal Institute of Technology from 1950 to 1954, and received the M.S. degree in 1954. Since that time he has been an assistant, and later Research Fellow, in the Department of Advanced Electrical Engineering, Swiss Federal Institute of Technology, Zürich, Switzerland.

B. Schneider

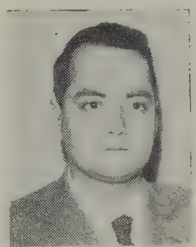
Mr. Schneider worked on a variety of transistor circuit problems, and has done research on junction diode and transistor noise. He has published several papers on temperature and other stabilization of transistor circuits, and on semiconductor noise.

❖

Max J. O. Strutt, for a photograph and biography please see page 925 of the May, 1958 issue of Proceedings.

❖

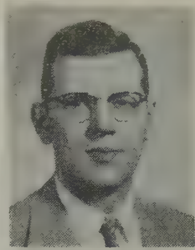
Jan B. van Erp was born in Laras, Simoeloengoen, Sumatra, on September 15, 1929. After graduating as an electrical engineer in January, 1955, from the Technological University of Delft, Netherlands, he worked several years as a research engineer in the measurement and control department of the Royal Dutch/Shell Laboratory at Delft.



J. B. van Erp

In January, 1958, he came to the United States to gain experience in the field of nuclear engineering. At present he is at the Argonne National Laboratory, Lemont, Ill., doing work on the kinetics, control, and instrumentation of nuclear reactors. Since January, 1959, Mr. Van Erp has been with "Euratom," Brussels, Belgium.

Lawrence J. Varnerin (SM'58) was born in Boston, Mass., on July 10, 1923. He received the B.S. degree in 1947 and the Ph.D. degree in 1949, in physics from the Massachusetts Institute of Technology.



L. J. Varnerin

From 1949 to 1952, he worked on deionization phenomena in gases at Sylvania's Electronics Division. From 1952 to 1957, he worked on surface physics, high vacuum studies and ionic pumping at the Westinghouse Research Laboratories. He then joined Bell Telephone Laboratories in 1957 where he has worked on gas discharges and high-frequency germanium transistors. He is presently concerned with components development.

Dr. Varnerin is a member of the American Physical Society.

❖

Ronald L. Wigington (M'57) was born in Topeka, Kan., on May 11, 1932. He received the B.S. degree in engineering physics from the University of Kansas, Lawrence, Kan., in 1953 and is currently attending graduate school at the University of Maryland, College Park, Md.

R. L. Wigington

From 1953 to 1954, he worked on millimeter wave problems in klystron exploratory development at the Bell Telephone Laboratories. While service on the Army from 1954 to 1956, he was assigned as an electronics engineer to the Department of Defense. Since that time he has been a civilian employee of the Department of Defense, working in the fields of millimicrosecond instrumentation and high-speed computation.

Mr. Wigington is a member of Tau Beta Pi, Sigma Pi Sigma, Pi Mu Epsilon, and is an associate member of Sigma Xi.

IRE Awards, 1959

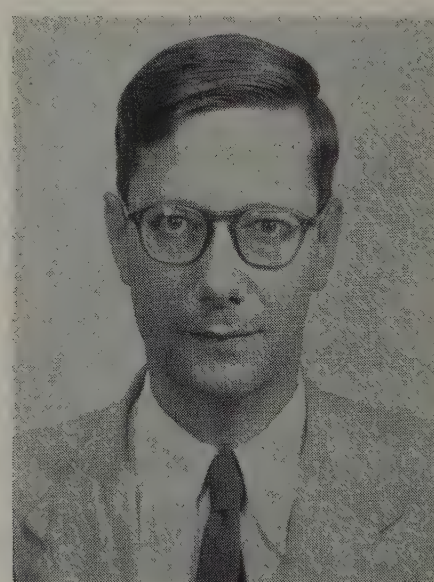
Medal of Honor Award

E. Leon Chaffee

For his outstanding research contributions and his dedication to training for leadership in radio engineering.

Joint Winners of the Morris Liebmann Memorial Prize

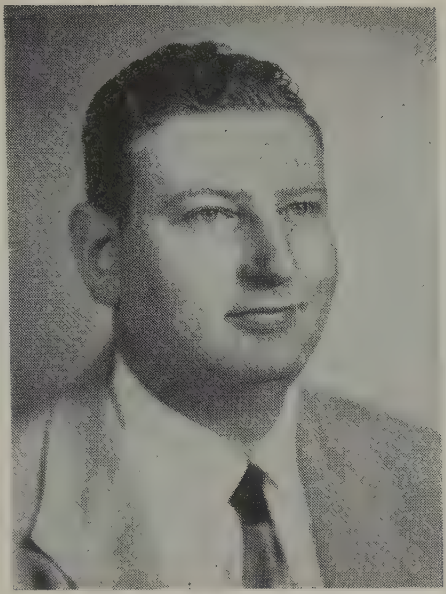
Charles H. Townes



Nicolaas Bloembergen

For important fundamental contributions to the maser.

Browder J. Thompson Memorial Prize



Franklin H. Blecher

For his paper entitled "Design Principles for Single Loop Transistor Feedback Amplifiers which appeared in the September, 1957 issue of the IRE Transactions on Circuit Theory.

W. R. G. Baker Award

Richard D. Thornton

For his paper entitled "Active RC Networks" which appeared in the September, 1957 issue of the IRE Transactions on Circuit Theory.

Vladimir K. Zworkin Television Prize

Paul Weimer

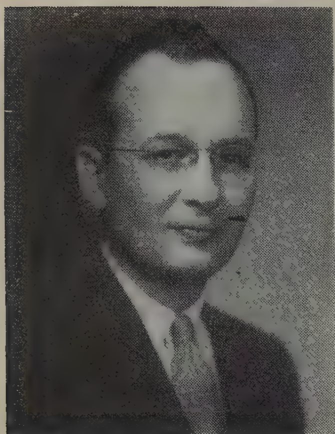
For contributions to photoconductive-type pickup tubes.

Harry Diamond Memorial Award

Jack W. Herbstreit

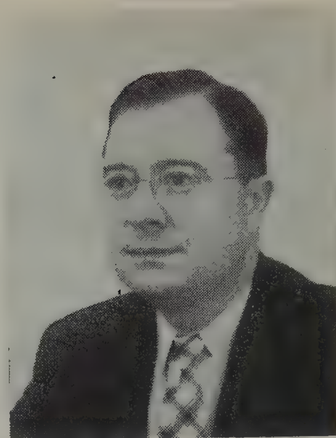
For original research and leadership in radio-wave propagation.

New Fellows



A. E. ANDERSON

For contributions to transistor switching circuits.

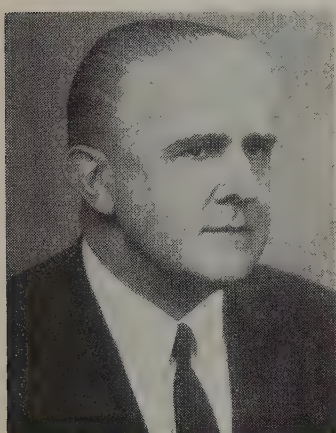


S. J. ANGELLO

For contributions in the field of solid-state rectifiers.

MAURICE APSTEIN

For contributions to the improvement of electronic ordnance devices.

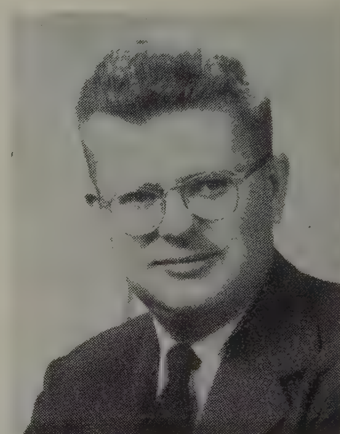


V. A. BABITS

For contributions to engineering education and pioneering in television.

J. M. BARSTOW

For contributions to the transmission of monochrome and color television.

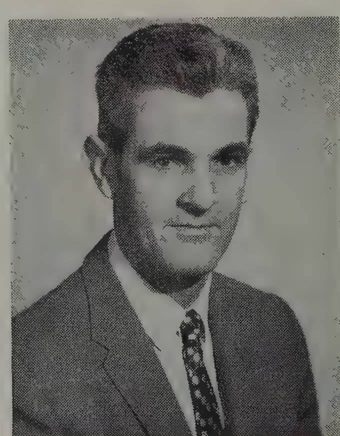


ROSS BATEMAN

For contributions in the field of ionospheric scatter propagation.

A. H. W. BECK

For contributions to the development of the thermionic valve.



J. T. BOLLJAHN

For contributions to the fundamental theory and design of antennas.

J. M. BRIDGES

For contributions to precision tracking radar and component reliability.

New Fellows

J. T. BROTHERS

For contributions and leadership in the field of electronic components.

W. C. BROWN

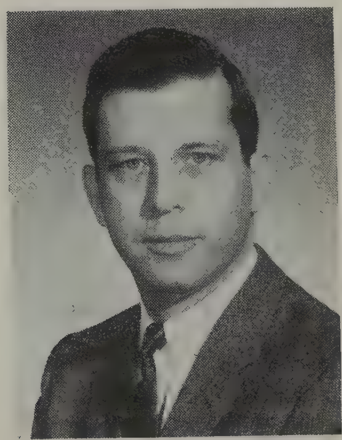
For contributions in the field of microwave tubes.

J. A. BRUSTMAN

For contributions in the development of digital computing systems.

C. C. CHAMBERS

For leadership in electrical engineering education.

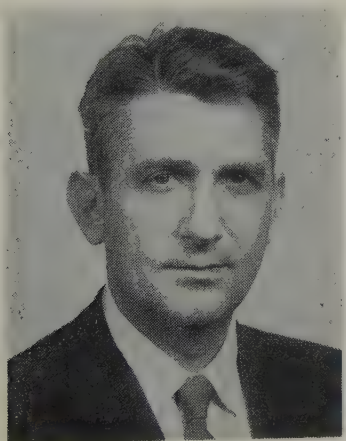


S. B. COHN

For contributions to the theory and design of microwave components.

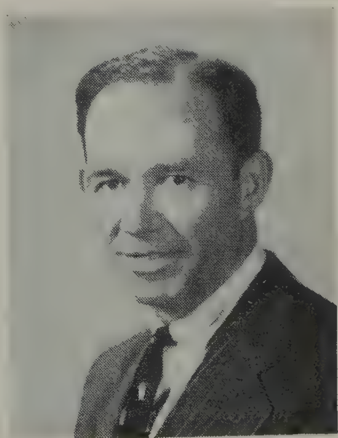
H. F. DART

For contributions to the electronic profession.



B. J. DASHER

For pioneering contributions in the field of network synthesis and engineering education.



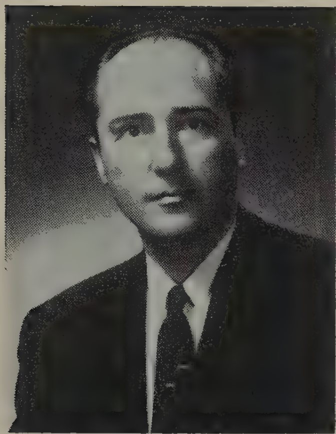
J. T. DEBETTENCOURT

For contributions to radio wave propagation theory and systems.

S. M. DEL CAMP

For contributions in the development and standardization of electronic components.

New Fellows

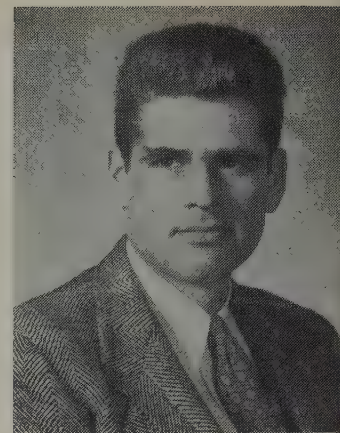


J. L. Dennis

For contributions and leadership in the field of aeronautical navigation.

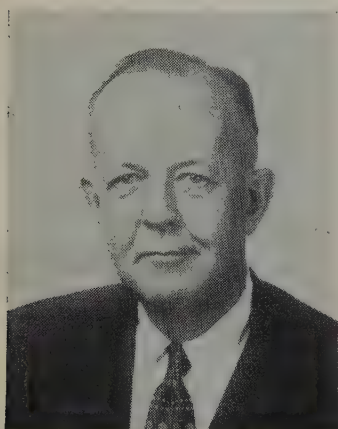
J. M. Early

For contributions in the development of high frequency transistors.



Peter Elias

For contributions to information theory and to engineering education.

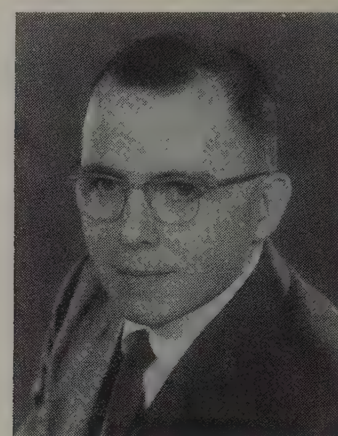


H. T. Engstrom

For contributions in the development and utilization of high-speed computers.

D. K. Gannett

For contributions to the transmission of television and sound broadcasting signals.

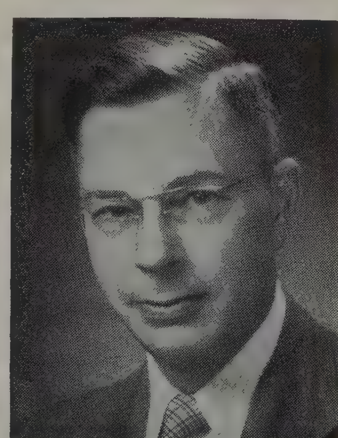


W. E. Good

For pioneering development in microwave spectroscopy and contributions to color television receivers.

R. E. Graham

For contributions in the field of radar tracking systems and television research.



George Grammer

For contributions in single sideband radio telephony and for publication of technical information.

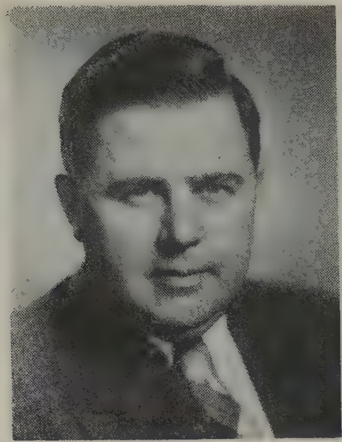
J. W. E. Griemsmann

For contributions to microwave research.

New Fellows

J. H. HAMMOND, JR.

For pioneering work in radio control and communication.



E. H. HANSEN

For contributions to the development of motion picture sound recording and reproduction.

E. F. HERZOG

For contributions to military electronics.

W. A. HIGINBOTHAM

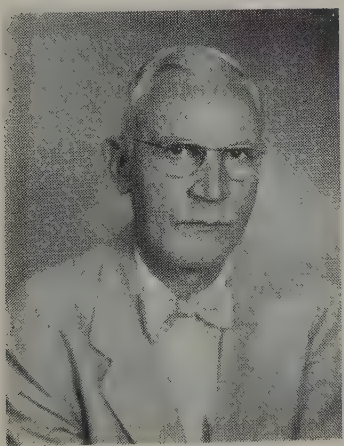
For contributions to pulse circuits and nuclear instrumentation.

L. N. HOLLAND

For contributions to engineering education.

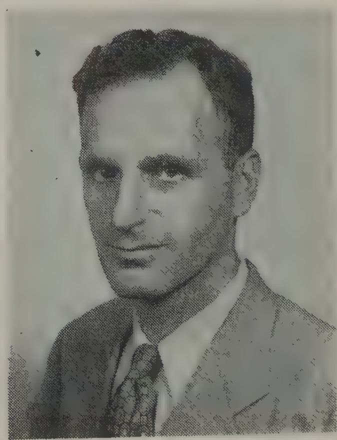
HEINZ HORN

For contributions in the development of ultra-high frequency and submarine cables.



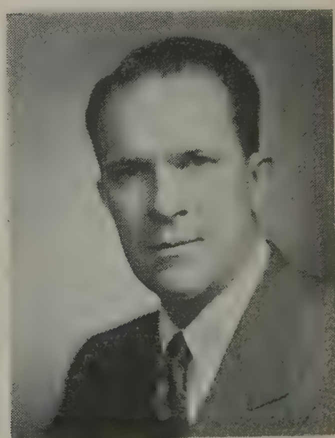
A. W. HORTON

For contributions to long distance telephony, electronic switching and military electronics.



H. E. KALLMANN

For contributions to transient response analysis of networks and to instrumentation.



M. V. KIEBERT, JR.

For contributions to telemetry and automatic data reduction.

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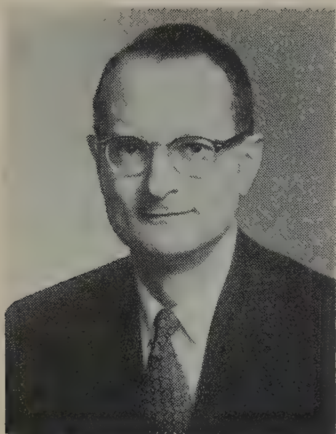


D. D. King

For contributions to microwave technology and electronic countermeasures.

T. P. Kinn

For contributions in the field of industrial heating.

L. K. Lee

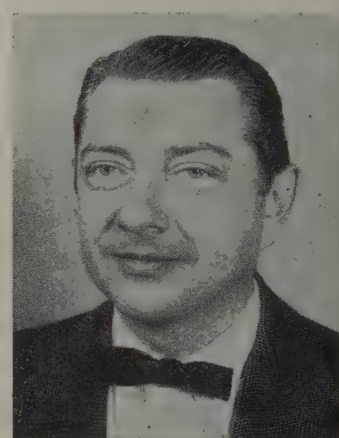
For contributions in the development of subminiature components and production techniques.

W. R. Le Page

For contributions to electronic engineering education and literature.

D. K. Lippincott
(deceased)

For outstanding service in the field of patent law.

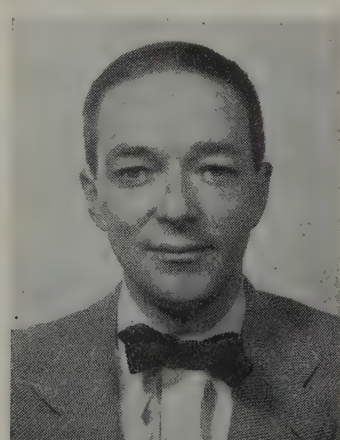


L. B. Lusted

For technical achievements and leadership in relating medicine and electronics.

J. R. Macdonald

For contributions to the theory of electronic circuits and electrical properties of solids.



S. J. Mason

For contributions in the fields of active networks and engineering education.

Frank Massa

For pioneering in electroacoustical engineering.

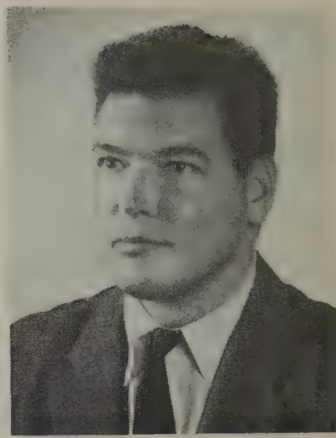
New Fellows

D. O. McCoy

For research in the field of radio astronomy.

R. D. McCoy

For contributions in the fields of servo-control systems and analog computers.



David Middleton

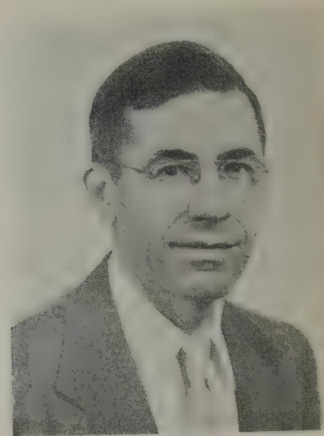
For contributions to the theory of noise in electronic systems.

Sidney Millman

For contributions in the fields of magnetrons, travelling-wave amplifiers, and backward-wave oscillators.

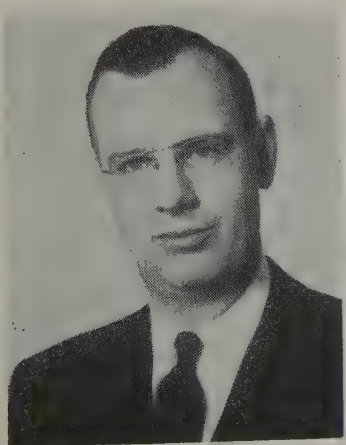
B. B. Minnium

For the development of electronic components and for service to the radio engineering profession.



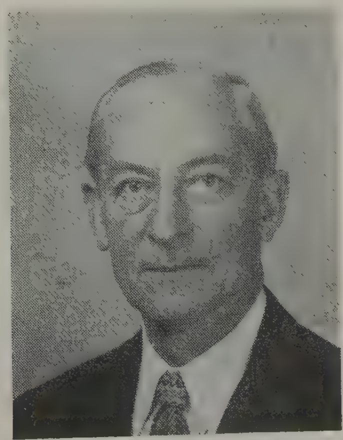
C. W. Mueller

For contributions to the development of electronic tubes and solid-state devices.



J. H. Mulligan, Jr.

For contributions in the fields of network theory, feedback systems, and engineering education.



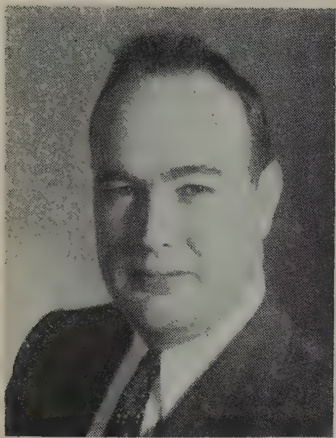
T. M. Odarenko

For contributions to the development of radio transmission systems, techniques, and components.

R. W. Olson

For leadership in geophysical research.

New Fellows

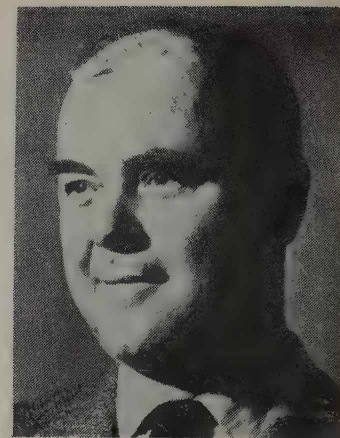


P. F. Ordung

For contributions to network theory and to electrical engineering education.

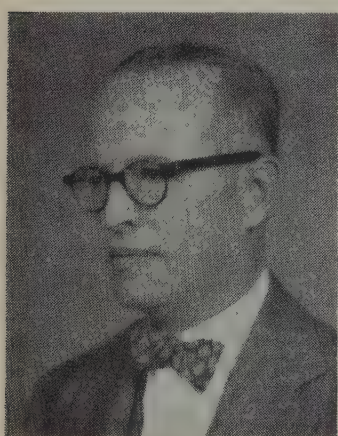
R. D. Parker

For pioneer developments of teleprinter and data transmission systems.



G. W. Patterson

For leadership in the development of machine logic and switching theory.

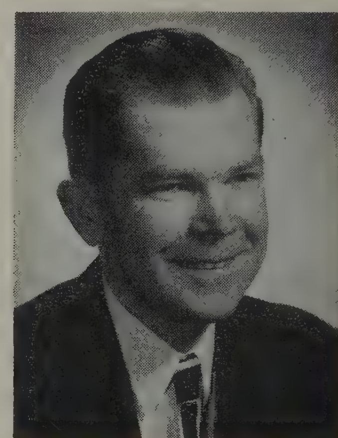


Eugene Peterson

For analysis and application of nonlinear devices in communication.

A. J. Pote

For contributions to high-power accelerators, radio-frequency heating, and communication systems.

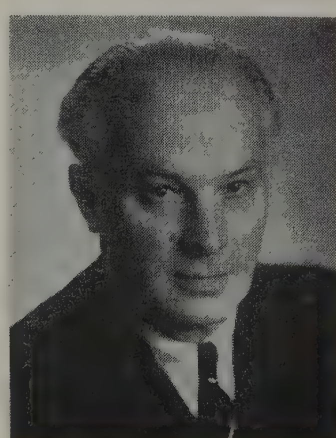


W. R. Rambo

For contributions to military electronics.

A. C. Rockwood

For contributions to standardization of electron devices.



Horst Rothe

For contributions in the fields of electron tubes and the theory of noise.

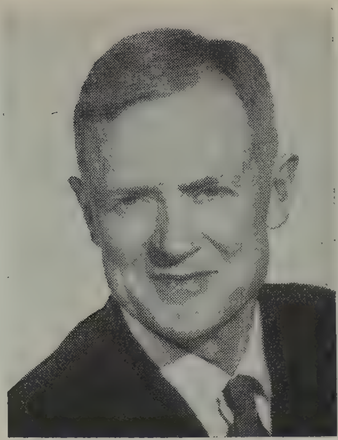
H. P. Schwan

For oustanding leadership in the medical electronic field.

New Fellows

L. C. SIMPSON

For contributions to broadcasting and telecommunication systems.



D. H. SLOAN

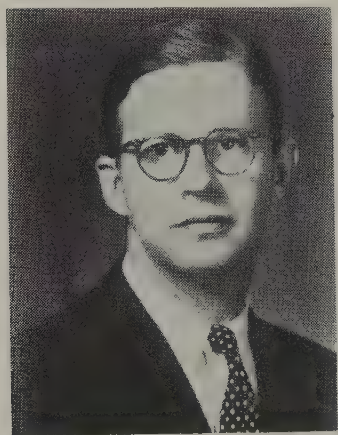
For contributions to high-power electron tubes.

O. J. M. SMITH

For contributions in the fields of electrical measurement and feedback control systems.

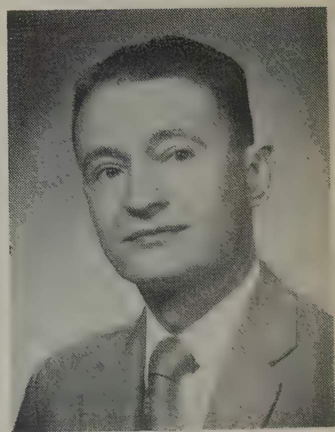
H. M. STEARNS

For contributions in the fields of microwave tubes and Doppler radar.



M. W. P. STRANDBERG

For research in the field of microwave spectroscopy.

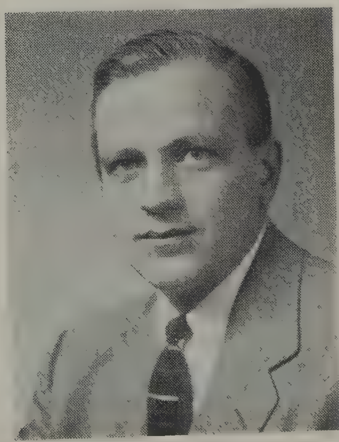


J. C. TELLIER

For contributions to receiver design and transistorization.

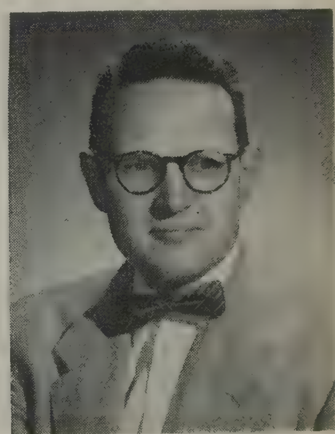
P. M. G. TOULON
(deceased)

For contributions to the control of gaseous conducting devices and in the field of color television.



J. G. TRUXAL

For fundamental contributions to the theory of feedback control systems.



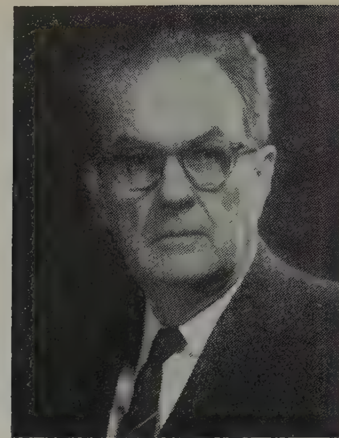
J. H. VOGELMAN

For contributions to military electronics.

New Fellows

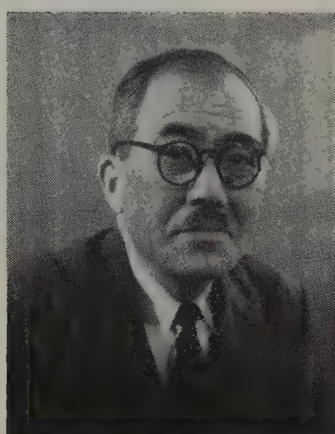
H. J. von Baeyer

For contributions to development of radio communication techniques and systems.



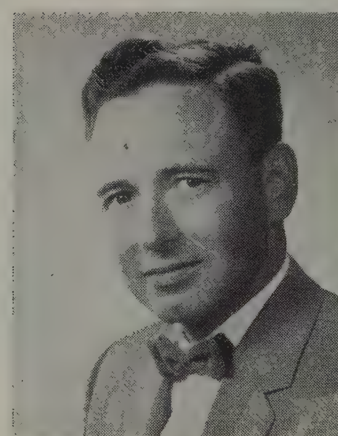
J. D. Wallace

For contributions in the fields of radio transmitters, antennas, and propagation.



Yasushi Watanabe

For contributions in the development of feedback amplifiers, transistor electronics, and to engineering education.



S. E. Webber

For contributions to high-power klystron amplifiers.

Scanning the Transactions

Synthesis by Computer. Designers of circuits and components are finding the digital computer an increasingly important tool for solving design formulas and for analyzing the performance of the resulting designs, especially when lengthy repetitive calculations are involved. In many cases, however, design formulas are only approximate and yield results which do not exactly match the desired performance specifications. The designer must then go the rest of the way toward the final design by trial and error methods, varying his parameters and analyzing the results over and over until the performance converges on the specification. Now it appears that this synthesis procedure, too, is being taken over by the computer. A computer was recently programmed not only to analyze each successive design, but also to decide which parameters to adjust next and by how much for the next trial run, so that entire cut-and-try process was carried out automatically in one rapid operation. Incidentally, the problem chosen for the experiment was the synthesis of a stagger-tuned three-cavity filter, which probably represented the first time a computer was used to solve a microwave circuit problem. (L. Young, "Microwave filter design using an electronic digital computer," IRE Trans. on Microwave Theory and Techniques, January, 1959.)

Broadcasters, too, are turning to computers. With new radio stations being added to the crowded broadcast band and existing stations trying to increase their power and improve their coverage, the problem of designing an antenna that will give good coverage without interfering with other stations is becoming increasingly intricate. Low-power stations are starting to use directional arrays with up to six towers—something considered uneconomical only a few years ago—to obtain the odd-shaped radiation patterns now required. And, if a station proposes to operate at night, sky-wave as well as ground-wave propagation must be taken into account. According to FCC rules this means that in addition to the horizontal radiation patterns, a complete azimuthal pattern must be computed for every five degrees of elevation up to sixty degrees. Thus not one, but 13 complex patterns have to be determined—a job ideally suited to the digital computer. (S. Bergen, "Calculation of directional antenna patterns using digital computer techniques," IRE Trans. on Broadcast Transmission Systems, January, 1959.)

Vitamin pills and railroad car wheels would have been the last things you would expect to find discussed in an IRE publication a half dozen years ago. But so rapidly have electronic techniques been adopted throughout science and industry that today almost any subject is fair game. Convincing evidence, if any is needed, may be found in the January issue of the IRE Transactions on Industrial Electronics. Vitamin pills? A photo-electric detection system is described which can accurately count and control the packaging of 15,000,000 capsules a day. Railroad car wheels? Positioning servos make possible a fully automated repair shop where old wheels are reconditioned and new ones are machined, bored, transferred and pressed on the axles—all automatically. What other jobs are being done electronically? Here is only a small sample: control of nuclear reactors, inspection of metal castings for structural flaws, reading 500 strain gauges at the rate of 50 per second, safety devices for preventing overspeed or overtorque of high-speed rotating machinery, and automatic machining of finished parts by numerical control systems.

Noise in stereo disk recordings, which is sometimes considered to be greater than in monaural recordings, has been found partly to originate from the geometry of the record playback system rather than from the recording channel itself. It arises from the fact that the pivot point about which the recording stylus moves is apt to be slightly different from the pivot for the playback stylus. This difference introduces a slight loss in playback level of the vertical component and some harmonic distortion. Moreover, because the pivot point is not in the plane of the record, a vertical motion of the stylus is accompanied by a horizontal component in the direction of the motion of the disk. This component alternately adds to or subtracts from the average groove velocity, resulting in frequency-modulation noise. These findings suggest that manufacturers can effect an improvement in the quality of stereo recordings by a careful and uniform selection of the vertical pivot point of tone arms. (D. Cronin, "Modulation noise in two-channel disk recordings," IRE Trans. on Audio, November–December, 1958.)

Hypodermic needles have never been a popular item with those on the receiving end. It now appears that those on the other end have had their problems, too. It seems that the victim's blood has a tendency to coagulate inside the needle, making it very difficult to clean. Ultrasonics, which has been used recently to clean everything from laundry to teeth to metal parts, has been found to provide a neat solution to the problem. Using ordinary tap water as the medium, an ultrasonic technique has been developed which not only removes dried blood quickly and effectively, but also makes it possible to sterilize the needle at the same time. While this development may be of small comfort to you in your hour of pain, it should be of substantial interest to hospital and laboratory technicians everywhere. (F. Nesh and J. R. Andreotti, "An ultrasonic cleaner for hypodermic needles and similar small bore apparatus," IRE Trans. on Ultrasonics Engineering, February, 1959.)

Megawatt traveling-wave tubes, predicted in a Proceedings paper three years ago, have now become a reality. At that time an experimental model had been operated at 300 kw pulsed power output, indicating that a megawatt tube was entirely feasible. Now this goal has actually been achieved, opening up a whole new range of power capabilities for the traveling-wave tube which will be of major interest to system designers as well as to microwave tube specialists. (M. Chodorow, *et al*, "The design and characteristics of a megawatt space-harmonic traveling-wave tube," IRE Trans. on Electron Devices, January, 1959.)

Two bibliographies of special interest were published in January. The first was contained in a report of an URSI Subcommission on antennas and waveguides. The report gives an authoritative summary of recent developments, primarily in the U. S., in some of the more active areas of antenna and diffraction theory, accompanied by 169 literature references. The second bibliography gives what is probably the most complete list yet to be compiled of papers dealing with masers and parametric amplifiers. The list of 121 references is appended to an excellent review of the methods of operation and the performance that has been achieved by these important new types of microwave amplifiers. (R. S. Elliott, *et al*, URSI Report on antennas and waveguides, and annotated bibliography," IRE Trans. on Antennas and Propagation, January, 1959; H. Heffner, "Solid-state microwave amplifiers," IRE Trans. on Microwave Theory and Techniques, January, 1959.)

Books

Gmelins Handbuch der Anorganischen Chemie: Germanium, 8th Ed. Rev., edited by E. H. Erich Pietsch.

Published (1958) by Verlag Chemie, GMBH, Weinheim/Bergstrasse, Germany. 576 pages+XLIV pages. 290 figs. $6\frac{7}{8}\times10$. DM337.—(In German)

The new volume of "Gmelins Handbuch" on germanium is a very exhaustive and detailed summary of all that is known about the physical and chemical properties of germanium and germanium compounds. Over 300 pages are devoted to electrical and optical properties of germanium, including a thorough coverage of *p-n* junctions, diodes, transistors, photocells, and other devices. The volume is indexed according to subject matter; the index is in German and English and is very complete, although there is no author index.

The treatments of the various subject headings are in the nature of review articles, with copious references to technical literature. The caliber of these reviews is generally high, and the coverage of the literature very inclusive. The book should prove very useful to investigators who wish an over-all view of a given topic and will also provide a guide to further study of the technical articles themselves.

Unfortunately, the survey of the literature is complete only up to 1955. This point limits the usefulness of the volume as a general literature survey, although the fact that the major effort of research and development work has shifted from germanium to other materials in the past few years makes this a less serious shortcoming than it might otherwise have been. The editors have been liberal in including a large number of illustrations and diagrams to accompany the text.

In summary, this edition of "Gmelins Handbuch" on germanium should provide the research worker or development engineer with a conveniently indexed, comprehensive, and reasonably complete review of what knowledge is currently available about germanium, germanium electron devices, and germanium compounds.

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Dynamical Analogies, 2nd edition, by Harry F. Olson

Published (1958) by D. Van Nostrand Co., Inc., 120 Alexander St., Princeton, N. J. 269 pages+8 index pages+xi pages. Illus. $6\frac{1}{4}\times9\frac{1}{4}$. $6.75.

H. F. Olson's "Dynamical Analogies" is a modification of the book of the same title first published in 1943. Except for minor changes, the first eleven chapters are essentially the same as the earlier edition. New chapters entitled, "Noise and Distortion," "Feedback," "Mobility Analogies," and "Mechanical Analogies" have been introduced. They constitute about 70 pages in addition to the 200 pages of the previous chapters.

The book as a whole presents in clear, readable form, with interesting diagrams, the extensive experience of the author in the application of analogies to the solution of dynamical problems in electroacoustics. The short chapter on noise and distortion gives formulas for calculating the noise due to thermal agitation of the air particles, the atoms in the vibrating system as well as the electrons in a conductor. The description of distortion in dynamical systems is short and nonanalytical.

The chapter on feedback should be quite useful. It assembles in one place, in systematic form, the dynamical formulas relating to the stabilization of a system that contains in it internal disturbance. Examples involving hydraulic regulator, an engine governor, power steering in automobiles, and an electronic feedback amplifier are discussed.

It is pleasant to see the addition of the chapter on mobility analogy. Many of us have made extensive use of this type of analogy in our work and writings and believe that it should be given emphasis in any book on this subject. The principal reference to this subject used by Olson is Floyd A. Firestone who has been the important leader in presenting the mobility concept to electroacoustic readers in this country.

The final chapter on the magnetic analogy rounds out the text. The material in this chapter, although not new to workers in the field, is an important part of the text as a whole.

The reviewer recommends this book to all who wish to apply the tool of electrical circuit theory to the solution of mechano-acoustic problems.

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Bolt, Beranek and Newman
Cambridge, Mass.

Sampled-Data Control Systems, by Eliahu I. Jury

Published (1958) by John Wiley and Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. 403 pages+5 index pages+6 bibliography pages+9 appendix pages+25 problems pages+xv pages. Illus. $6\times9\frac{1}{4}$. $16.00.

The theory and practice of sampled-data control systems have been developing rapidly during recent years, and Dr. Jury, who has contributed heavily to this development, has written an excellent book.

This reviewer has been particularly impressed by the wealth of subjects that have been more than adequately covered. Written for the graduate student and practicing engineer who have mastered the fundamentals of feedback system theory and the theory of complex variables, the purpose of the book as outlined by the author is "—to develop a basic theory that can be applied to sampled-data systems, to other allied fields such as circuits, networks and computers, and to the general field of systems engineering." As is to be expected in any book dealing with control systems, careful attention is given to the problems of analysis and synthesis, with five chapters concerned with the former and three with the latter.

In the first chapter the concept of a sampled-data system is introduced and the mathematical description of the sampling process is given. The z-transform method is defined, and its relationship to the Laplace transform is shown clearly. With these fundamentals established, other useful concepts are developed. These include the sampled-data system transfer function, inverse z-transform, transformation theorems and stability. As in the remainder of the book, the presentations include the supporting mathematics. The chapter concludes by examining the characteristics of hold circuits. A fairly complete table of z-transform pairs is given also.

As the z-transform method is limited to yielding results at the sampling instants only, the second chapter introduces the modified z-transform which removes this restriction. The same subjects investigated for the z-transform are presented from the point of view of the modified z-transform approach. Tables of modified z-transforms are given together with a table showing the transfer functions for various feedback loop configurations.

Chapter three is devoted to the root-locus method of analysis of sampled-data systems. An illustration of the use of the root-locus in system compensation is given in some detail. The remaining material in this chapter covers constant overshoot loci, equations for overshoot and peak time and the correlation between frequency-locus and root-locus.

In chapter four, standard frequency response methods of analysis are extended to sampled-data systems. The relationship of the sampled-data frequency spectrum to the continuous frequency spectrum is presented. Once this is done, the techniques of the Nyquist diagram and Bode plots follow in a logical manner.

Chapters five through seven take up the problems of synthesis. Chapter five deals with discrete compensation methods, *i.e.*, the use of digital computers, or a number of delay elements. Three interesting examples are worked in detail. Chapter six discusses system compensation through the use of continuous networks. The synthesis portion of the book concludes in chapter seven which goes into the physical realization of the discrete compensators. There is enough material presented in the synthesis chapters to gain an appreciation for the practical difficulties that occur. Facing a real problem, the designer should have obtained enough insight to choose between the digital and continuous compensation methods; the one having extreme flexibility, the other usually much simpler in equipment requirements.

Chapter eight presents some of the methods for analyzing continuous systems by sampled-data techniques. Included are the fictitious hold method and the z-form and

modified z-form methods. The chapter concludes by applying the method of z-transforms to the solution of difference equations with constant and periodic coefficients.

The final chapter takes a closer look at the methods of sampling analysis by presenting a clear picture of what is involved in the assumption that the sampled signal can be treated as a train of impulses. The P-transform method is then derived to be used in systems in which the sampling pulse width is not negligible in comparison to the system time constant.

All in all, the book presents the sampled-data picture remarkably well and should be welcomed by those desiring to learn about this subject, and by those who desire a comprehensive reference.

In addition, the excellence of the choice of problems and of the many drawings should be mentioned.

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Handbook of Automation, Computation, and Control, Vol. 1: Control Fundamentals, edited by E. M. Grabbe, S. Ramo, and D. E. Wooldridge

Published (1958) by John Wiley and Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. 974 pages+20 index pages+xx pages. Illus. $6\frac{1}{4} \times 9\frac{1}{4}$. $17.00.

"Control Fundamentals" is the first of a set of three volumes whose objective is to give "problem solvers" in the fields of automation, computation, and control a condensed presentation of "the available theory and information on general mathematics, feedback control, computers, data processing, and system design." Volumes 2 and 3, which have not appeared yet, will cover "Computers and Data Processing" and "Systems and Components."

To a gratifying extent the editors and the twenty-nine contributors of Volume 1 succeed in covering a great deal of the basic knowledge required by control systems engineers. The following are some of the chapter headings (there are twenty-four chapters in all): "Sets and Relations," "Matrix Theory," "Finite Difference Equations," "Operational Mathematics," "Conformal Mapping," "Boolean Algebra," "Statistics," "Numerical Analysis," "Operations Research" (which includes discussions of game theory, allocation models, waiting time models, etc.), "Information Theory," "Relation Between Transient and Frequency Response," "Feedback System Compensation," "Noise, Random Inputs and Extraneous Signals," "Nonlinear Systems," and "Sampled-Data Systems and Periodic Controllers."

Each chapter is essentially a condensation of most of the fundamental results of a special body of theory, and for the most part good lists of references are provided. As a result the user of the "handbook" will find that it can be a good point of departure for many of his present needs in the realm of basic theory, if it doesn't already provide him with all he wants to know. The minimum previous training he will need is included in the technical portion of most undergraduate engineering curricula.

In a work of this kind, where the subject matter is so broad, it is unavoidable that many users will not find sufficient variety or depth of material. The editors have to stop somewhere, of course. We suggest, however, that certain additional subjects could be usefully included, perhaps in a supplementary volume. The following areas, in this reviewer's opinion, deserve such treatment: calculus of variations, dynamic programming, analysis of automata and neural networks, classical mechanics (with emphasis on the ballistics of space missiles), theories of self-optimizing and adaptive systems, multidimensional control systems, final-value control systems, and aspects of human psychology involved in matching control systems to their users and operators.

These lacks, however, do not detract from the value of the huge amount of basic information that has already been packed into this volume. We believe it will serve as an excellent reference work.

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Princeton, N. J.

Nonlinear Problems in Random Theory, by Norbert Weiner

Published (1958) by The Technology Press, M.I.T., Cambridge, Mass., and John Wiley and Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. 128 pages+3 index pages+ix pages. Illus. $6 \times 9\frac{1}{4}$. $4.50.

This interesting monograph represents some of Professor Wiener's most recent work in random theory. Here, he is particularly concerned with nonlinear problems in a variety of fields. A brief review of the topics discussed will give the reader an idea of the scope of his interests. The book itself consists of 15 lectures which were recorded and subsequently reproduced in mimeographed notes. These lectures contain a description of a class of Gaussian random processes, in particular, the Brownian Motion processes and their representations. Considerable attention is given to a treatment of both linear and nonlinear functionals of such processes. After four preliminary lectures, Professor Wiener applies his methods first to a discussion of frequency modulation. Two types of frequency-modulation problems are considered. Type 1 is a linear frequency modulation where the random variations in frequency appear as a linear functional of the driving stochastic mechanism. Type 2 imparts a quadratic stochastic driving mechanism. The result of suitable operations then yields the desired spectral density of the frequency-modulation process. This in turn is applied to a study of brain waves, and "random time," in lecture 8. In lecture 9, some questions of using Brownian processes to describe probabilities in quantum theory are considered. In lectures 10 and 11 Professor Wiener next applies these ideas to still another area of interest, specifically the problem of analyzing and synthesizing nonlinear electrical systems. Here a typical problem is to determine the operational characteristics of an unknown linear or nonlinear "black box." In lectures 12 and 13 his methods are applied to coding and decoding problems. Finally, in lectures 14 and 15, some new approaches to the description of statistical mechanical systems, such as a three-dimensional gas subject to various types of force fields, are outlined.

While this book is not directed towards the average engineer there is probably no technical reader who will not enjoy and profit from some of the discussions. Those who will find it most useful, perhaps, are the communication engineers, although physicists and engineers with an interest in hydrodynamics, statistical mechanics, and quantum mechanics will find interesting ideas in the various lectures. Nonlinear problems occur, of course, in all the areas mentioned. Moreover, they have received considerable attention in recent years. Although the present book does give a powerful approach to such problems, with much new material, it should be pointed out that considerable progress has also been made along similar lines by a number of other investigators in the last decade. For example, similar questions involving nonlinear functionals, the distributions of noise processes after nonlinear devices, and studies of frequency modulation, have been considered by Kac, Siegert, Darling and others, and in the case of frequency modulation by Middleton, to mention but a few.[1] This excellent introduction to the subject which Professor Wiener has given us would be enhanced still further if a more comprehensive bibliography and reference to those and related studies had been made available to the reader.

In summary, we are indebted to Professor Wiener for a glimpse into new problems and techniques involving random processes. His monograph is timely, with a delightfully informal style, depending much on the way it was originally recorded and a vigorous approach, which suggests new discoveries to come. [A number of typographical errors are easily corrected by the reader, (cf. Eqs. (4.13), (5.12), (5.34), (6.9), (6.15), (6.16), (6.26), (6.30).)]

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[1] See, for example, M. Kac and A. J. F. Siegert, "On the theory of noise in radio receivers with square law detectors," *J. App. Phys.* vol. 18, p. 383; (1957).
A. J. F. Siegert, "Passage of stationary processes through linear and nonlinear devices," IRE TRANS. ON INFORMATION THEORY, vol. 3, p. 4; March, 1954.
D. A. Darling and A. J. F. Siegert, "A systematic approach to a class of problems in the theory of noise and other random phenomena—part I," IRE TRANS. ON INFORMATION THEORY, vol. 3, p. 32; March, 1957.
A. J. F. Siegert, "Part II, Examples," *ibid.*
D. Middleton, "The distribution of energy in randomly modulated waves," *Phil. Mag.* ser. 7, vol. 42, p. 689; 1951.

Fundamentals of Advanced Missiles by Richard B. Dow

Published (1958) by John Wiley and Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. 555 pages+11 index pages+xvi pages. Illus. $6 \times 9\frac{1}{4}$. $11.75.

This book represents an excellent introduction to the field of guided missilery. It is indeed a summary, but it makes a success of an attempt to relate fundamental principles and their applications to the major technical areas in which missile designers must be proficient. The term "advanced" in the title is misleading unless one considers that all missiles are "advanced." The text is really at an introductory level but makes a welcome addition to the library of specialists who want a broad look at missile problems

or to the beginner who needs to know some of the problems ahead.

The author has contrived to gather an excellent summary of a broad field and to bridge the gap between fundamentals and applications by a careful arrangement of topics. For example, principles of propulsion and properties of fuels are discussed in the chapter on fluid mechanics; forces on the airframe, dynamic stability, and automatic control systems follow the introductory material on dynamics; reliability, signal to noise problems, and kill probability are included in the chapter on probability and statistics; and properties of microwaves, wave propagation, circuit elements such as magnetrons and klystrons, beam characteristics, and infrared radiation are discussed as a prelude to radar and guidance chapters.

The result is a logical presentation of most of the subsystem elements required to make up the airborne portion of a guided missile weapon system.

The chapter on systems is lacking in attention to a discussion of exchange or tradeoff ratios, growth factors, subsystem integration, and the broader implications and scope of systems engineering but does present several typical system analyses.

The extensive use of footnotes adds much useful information without detracting from the readability of the book. The Table of Contents is sufficiently detailed to be of major assistance in locating material in general and the index is moderately well done. The author is to be commended for this addition to the literature on guided missile design.

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Microwave Propagation in Snowy Districts, edited by Y. Asami

Published (1958) by the Research Institute of Applied Electricity, Hokkaido University, Sapporo, Japan. 198 pages. Illus. $7\frac{3}{8} \times 10\frac{1}{4}$.

This book can be of great value to any engineer engaged in establishing or maintaining a communication system in a region where snow and ice are listed among the facts of life. The only fault I have to find with the book is its title. The book does deal with propagation through snow and over snow-covered terrain, but it is also concerned with the accretion of snow and ice on antennas. Y. Asami, editor, and coauthor of two of the eight papers in the book, is Chairman of a committee of the Institute of Electrical Communication Engineers of Japan, which has been studying the problems involved in microwave communication in snowy districts. As a result, a book has been produced which examines these problems and their solutions with great thoroughness, both experimentally and theoretically.

The book contains eight papers by a total of nine authors and co-authors. The substance of the book is to be found in the first five papers. The first two cover the mechanism of ice-formation on antennas, the physical properties of this ice, and dielectric properties of many different forms of snow and ice, and the variation of these properties with frequency, temperature, and time. In the third paper the effect of ice and snow accretion on antenna performance, both as regards attenuation and antenna matching, is considered experimentally and theoretically. Then, in logical order, the problem of anti-icing and de-icing by electric heating is discussed. The fifth paper deals with the reflection coefficient of snow-covered terrain and the formation of atmospheric ducts over such a surface.

The subjects of the last three papers are the statistical examination of propagation data and a theory of space diversity. Although these papers are interesting in themselves, they do not fall within the main theme, and occupy only 36 of the 198 pages.

The papers are of the nature of comprehensive summaries of the present state of knowledge in the field, and yet contain much information which is original to the Japanese authors. The book is to be highly recommended to anyone faced with the task of communication in snowy districts.

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Natl. Res. Council of Can.
Ottawa, Ontario, Canada

Junction Transistor Electronics, by Richard B. Hurley

Published (1958) by John Wiley and Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. 461 pages+11 index pages+xvii pages. Illus. $6 \times 9\frac{1}{4}$. \$12.50.

The term "transistor electronics" has been used to cover a wide range of subject matter, from the physics of junction-transistor operation to a description of transistor applications. Mr. Hurley's book is concerned with fundamentals of transistor circuits and is intended for the circuit man. According to the author, the text is aimed primarily at the electronics engineer who has had first courses in vacuum-tube electronics and electric circuits.

The first two chapters, on physics of semiconductors and of transistor operation, are intended to be more descriptive than rigorous. Accordingly, these chapters—along with similar material in later portions of the book—should be approached with some caution. For example, in Chapter 2, the author speaks of "forward carriers diffuse through a depletion layer." In Chapter 8, in discussing high-level operation of a *p-n-p* transistor, when a large number of holes is injected into the base from the emitter, the author states that charge neutrality is preserved by causing an "equal number of electrons to be drawn in from the emitter" (rather than from the base contact).

At the end of the second chapter, the author derives an equivalent circuit for the transistor based on its physical operation. Except for the criticism noted above, this reviewer is glad to see this approach. Thus, at the outset, the student is aware of the physical phenomena taking place in the junction transistor, rather than just being given an equivalent circuit with numbers to be obtained from a manufacturer's data sheet.

Subsequently, the physically derived equivalent circuit is converted to the usual equivalent-Tee circuit, and Chapter 3 contains a tabulation of gain and impedance equations in terms of the Tee parameters. The remainder of the first half of the book comprises chapters on audio-frequency, amplifiers, power amplifiers, and dc amplifiers, biasing, noise, feedback, and ends with an interesting chapter on regulated power supplies.

In some respects, a different arrangement of material might have been desirable. For example, as noted above, junction-transistor operation is described in Chapter 2, alone with derivation of an equivalent circuit, but characteristic curves are not discussed until six chapters later. Similarly, biasing schemes are discussed in Chapter 5, following a discussion of the operation of audio-amplifiers in Chapter 4. Actually, the author had anticipated these criticisms because in the Preface he indicates that his arrangement of material and treatment is based on his own experience in the field and by the reactions of his students.

Explanations of transistor circuits are facilitated by liberal use of numerical examples. Polarities and magnitudes of voltages across the terminals of transistors and other components are indicated in many circuit diagrams; this is quite helpful to the student. Each chapter has a dozen or more references. Some of these, however, refer to material of limited availability (*e.g.*, company reports), which occasionally is somewhat frustrating to a reader.

The second half of the book begins with a chapter on high-frequency equivalent circuits. Most of these, in this reviewer's opinion, are too complicated for practical circuit use. Moreover, some of the more widely used equivalent circuits, *e.g.*, Giacoletto's hybrid-Pi, are not discussed. Subsequent chapters are devoted to internal feedback, video and tuned amplifiers, oscillators, and modulators. In general, the material here is not presented in as much detail as in the first half of the book. However, the style of the first half of the book returns in Chapters 19–21, which are devoted to switching circuits. Also included here is a discussion of binary artihmetic and logic circuits. The final chapter is devoted to the subject of saturable reactor circuits.

In summary, then, this book is a text rather than a reference work. Its principal applications probably will be in teaching a circuit man the fundamentals of transistor circuits—but not transistor physics.

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Handbook of Physics, edited by E. U. Condon and Hugh Odishaw

Published (1958) by McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. 1402 pages+10 index pages+13 appendix pages+xxvi pages. Illus. $7\frac{1}{4} \times 10$. \$25.00.

The "Handbook of Physics" gathers in one volume much of the detailed information essential to a comprehensive knowledge of physics. Prepared by a staff of about ninety specialists, this handbook is intended as a judicious selection from the vast literature of physics of "what every physicist should know."

In contrast to the "Handbook of Chemistry and Physics" and the "American Institute of Physics Handbook," for example, the present volume is not concerned

th exhaustive compilations of widely nging experimental data and practical chnological information. Unlike the re- ntly published "Fundamental Formulas Physics" (an excellent reference work in s own right), the "Handbook of Physics" aces emphasis less on the listing of formu- s, and more on a consideration of formulas their proper perspective.

In effect, it is a carefully organized col- ction of abridged textbooks covering every ajor branch of physics. Its scope is perhaps est indicated by a brief survey of its prin- pal parts: mathematics; mechanics of articles and rigid bodies; mechanics of de- rmable bodies; electricity and magnetism; eat and thermodynamics; optics; atomic hysics; the solid state; and nuclear physics. or the most part, each chapter is the coun- erpart of a full-length textbook or treatise. he chapters on electricity and magnetism, or example, are as follows: basic electro- nagnetic phenomena; static electric and nagnetic fields; electric circuits; electronic ircuits; electrical measurements; conduc- ion; metals and semiconductors; dielectrics; nagnetic materials; electrolytic conductiv- ty and electrode processes; and conduction f electricity in gases.

Since most of the chapters are written n a level which should be found compre- ensible to a nonspecialist, the student is likely to find the Handbook useful not only as a reference work, but also as a means of gaining entry into a new field. In most chap- ters, the lay of the land is sketched out in bold strokes, and usually there is ample bibliographical material to guide further pursuit. The working scientist is likely to find the Handbook more convenient to have on his desk than an odd assortment of un- dergraduate and possibly graduate text- books, particularly because it is well indexed, and because it contains a very useful appen- dix on units and conversion factors. For ad- vanced work, of course, the Handbook is no replacement for the more than fifty volume "Encyclopedia of Physics" (Handbuch der Physik).

The large size and weight (5.8 pounds) of the Handbook pose a practical handling problem, but this minor disadvantage is off- set by its overall convenience. This has been verified in practice by three months use on this reviewer's desk.

The editors, E. U. Condon and H. Odi- shaw, have exercised a great deal of wisdom in their choice of subject matter, though each reader is likely to wish that certain subjects had been treated in greater detail. For instance, this reader would have pre- ferred a more thorough coverage of mathe- matical physics, possibly at the expense of other topics. A fair question is: Why is group theory, a branch of mathematics widely used in physics, treated in little more than one page, when thirty-one pages are devoted to an article entitled "Fundamental Constants of Atomic Physics," which can be found in substantially the same form elsewhere? From the standpoint of gaining a liberal education in physics, the general reader is more likely to profit from reading this article than a comparable article on group theory. However, the balance would appear to be in favor of the group theory article in a refer- ence work.

This volume is naturally at its best on the more elementary subjects, and on those least sensitive to recent developments. A few of the advanced chapters, which at- tempt to summarize the status of a rapidly changing field (such as microwave spectros- copy), appear slightly out of date.

In summary, the "Handbook of Physics" is well written, well organized, and well in- dexed, and as such is a welcome addition to the reference literature. It is also beauti- fully printed in a pleasing typography. It is highly recommended to students and work- ing scientists in all branches of science and technology. It makes a nice gift for an aspir- ing scientist.

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Abstracts of IRE Transactions

The following issues of Transactions have recently been published, and are now available from the Institute of Radio Engineers, Inc., 1 East 79th Street, New York 21, N. Y. at the following prices. The contents of each issue and, where available, abstracts of technical papers are given below.

Sponsoring Group	Publication	Group Members	IRE Members	Non-Members*
Antennas & Propagation	Vol. AP-7, No. 1	$2.50	$3.75	$7.50
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Antennas and Propagation

Vol. AP-7, No. 1, January, 1959

Propagation of a Ground Wave Pulse Around a Finitely Conducting Spherical Earth from a Damped Sinusoidal Source Current—J. R. Johler and L. C. Walters (p. 1)

The form of the transient electromagnetic ground wave which has been propagated over a finitely conducting spherical earth from a source current dipole can be calculated by a direct quadrature evaluation of the Fourier integral. The method is illustrated in this paper by a calculation of the transient field radiated by the particular case of the damped sinusoidal source current dipole. At short distances from the source, the earth was assumed to be a plane and the displacement currents in the earth were neglected. The pulse was then calcu- lated by a direct evaluation of the Fourier integral and the integration was verified by special operational methods (inverse Laplace transformation). The form of this pulse was then predicted at great distance from the source by a direct evaluation of the Fourier integral in which the displacement currents in the earth and the earth's curvature were introduced into the Fourier transform. The form of the tran-

sient signal was found to be dispersed by the propagation medium. The most noteworthy attribute of this dispersion is a stretching of the period of the wave so that the form of the source is somewhat obscured by the filtering action of the medium.

On the Measurement of Virtual Height—I. Kay (p. 11)

A time dependent definition of the virtual height of a reflected wave train is suggested. This definition is such that its accuracy increases as the width of the incident pulse increases. Moreover, it is theoretically possible to obtain an estimate of the virtual height with as small an error as desired, no matter what the nature of the reflecting medium.

Suppose an incident pulse having width W and a carrier frequency ω_0 produces a reflected wave which, when measured at a fixed point $x=0$ in space, is $R(\omega_0, w, t)$. The suggested definition of the virtual height $h'(\omega_0)$ is

$$h'(\omega_0) = \lim_{w\to\infty} c\left\{ \int_{-\infty}^{\infty} t\,|R(\omega_0, w, t)|^2 dt \Big/ \int_{-\infty}^{\infty} |R(\omega_0, w, t)|^2 dt - \int_{-\infty}^{\infty} t\,|I(\omega_0, w, t)|^2 dt \Big/ \int_{-\infty}^{\infty} |I(\omega_0, w, t)|^2 dt \right\},$$

where c is the free space velocity of light. This relation for $h'(\omega_0)$ holds for any physically reasonable incident wave train and is independent of its envelope shape.

An example is given of a reflected wave whose virtual height cannot be determined by inspection in the usual manner. The expression given here for the virtual height provides the correct result.

Back-Scattering Measurements with a Space-Separation Method—H. J. Schmitt (p. 15)

A method for the experimental determination of the back-scattering cross section of arbitrarily shaped obstacles is suggested, which, in a manner analogous to the Michelson interferometer in optics, makes use of a semitransparent mirror in order to separate the incident wave and the reflected wave. A measurement setup is described, and possible sources of error are discussed. The accuracy of measurements is investigated by comparing the measured values of the back-scattering cross section of circular metallic disks with the results obtained from the exact theory.

Scattering of a Surface Wave by a Discontinuity in Reactance—Alan F. Kay (p. 22)

The following two-dimensional scattering problem is solved exactly by a Wiener-Hopf procedure. The incident field is a TM surface wave traveling in the positive x direction and guided by a reactive surface in the plane $z=0$. The surface has normal reactance X_0 if $x<0$, and X_1 if $x>0$. X_0 and X_1 are assumed positive and real. The discontinuity produces reflected and transmitted surface waves and a radiated field. Closed form expressions are found for the magnitudes of these fields. The reflected, radiated, and transmitted power flows, relative to that of the incident field, are plotted in universal curves. Conservation of energy is verified exactly.

Spherically Symmetric Lenses—Alan F. Kay (p. 32)

A design procedure is given for finding the index variation of a spherically (or cylindrically) symmetric lens which will produce, with some restrictions, any desired shaped beam pattern. Applications are made to a broad beam Luneberg lens, a bistatic Luneberg reflector, and a bistatic Eaton-Lippmann lens, and other examples are worked out. The practical importance of the index singularity at the center of lenses of the Eaton-Lippmann type is treated.

On the Design of Some Rhombic Antenna Arrays—A. A. de Carvalho Fernandes (p. 39)

The expression of the field radiated by a rhombic antenna, taking into consideration both the vertical and the horizontal polarization components, is used to establish the theory of the array of two stacked rhombics and of the array of four rhombics in a stacked and interlaced arrangement. The main conclusions obtained are that for convenient values of the vertical and horizontal spacings between the antennas of the array, there is a greater concentration of power radiated along the directions of the main lobe of the pattern and, as a result, these arrays show an appreciable gain over a conventional rhombic. Practical rules for the design of these arrays for point-to-point and broadcasting are given in some detail.

Radiation Field of an Elliptical Helical Antenna—J. Y. Wong and S. C. Loh (p. 46)

Rigorous expressions for the radiation field of a helical antenna of elliptical shape are derived on the assumption of a traveling-wave type of current distribution along the helix conductor. The analysis is valid for integral and nonintegral numbers of turns. These expressions for the general helix are employed to determine the fields for the limiting case of a circular helical antenna, and the results are essentially the same as those derived by both Knudsen and Kornhauser. Allowing the ellipse to degenerate to the other limiting case, a solution for the radiation field is obtained for the planar or commonly known zig-zag antenna. Therefore, it is possible to achieve a truly circularly polarized field with an elliptical helix of slight ellipticity.

The Rectangular Loop Antenna as a Dipole—R. W. P. King (p. 53)

An integral equation for the current in a rectangular loop of wire is derived for a loop that is driven by two generators located at the centers of one pair of opposite sides. The EMF's are equal in magnitude and in phase in the sense that they maintain currents in the generators that are in the same direction relative to the coordinate system and, therefore, in opposite directions from the point of view of circulation around the loop. An approximate solution is obtained for the distribution of current around the loop and for the driving-point impedance. It is shown that the solution for the rectangle of wire reduces to that of the symmetrically driven folded dipole when one dimension is made electrically small and to a section of transmission line driven simultaneously at both ends when the other dimension is made small. The loop that is electrically small in both directions is also examined.

Properties of Slotted Dielectric Interfaces—R. E. Collin (p. 62)

A theoretical analysis of slotted dielectric interfaces based on an application of the Rayleigh-Ritz method is presented. Formulas for calculating the equivalent circuit parameters are derived for arbitrary polarization and angles of incidence. Numerical results are given which show that the slotted dielectric interface behaves essentially as a homogeneous anisotropic dielectric interface. Formulas and numerical values of the equivalent dielectric constants are also given.

Traveling-Wave Cylindrical Antenna Design—A Graphical Synthesis Method—Peter Folds (p. 74)

A simple graphical method is given for the synthesis of a special line source system. The elements of the line source system are excited with equal amplitudes and continuously increasing phases. The sources lie on a cylindrical surface. The shape of this surface depends on the given pattern functions.

Theoretical Research on Tropospheric Scatter Propagation in the United States 1954–1957—H. Staras and A. D. Wheelon (p. 80)

This paper outlines recent progress in the theory of tropospheric scatter propagation in the United States. In the past three years, the emphasis of theoretical research has shifted from the analysis of the average signal level to the analysis of the signal statistics and to the underlying hydrodynamics of atmospheric turbulence. As might be expected in such a new and complex field, there is far from unanimity of opinion as to the "best" model to explain the myriad experimental results.

URSI Report on Antennas and Waveguides, and Annotated Bibliography—H. V. Cottony, *et al.* (p. 87)

Recent developments in some of the more active areas of antenna and diffraction theory are summarized. Although concerned primarily with U.S.A. activity, some reference is made to the literature of other countries where closely allied work is in progress. Fields included in this survey are broad-band antennas, wide-angle microwave optics, antennas for ionospheric scatter propagation, traveling-wave antennas, slot radiators, diffraction, and scattering theory.

Recent work in broad-band antennas has been concerned mainly with trying to hold *both* radiation pattern and impedance independent of frequency. The problem has been approached successfully through development of that class of antenna shapes which depends on angles only. The equiangular spiral is a simple example.

In microwave optics the requirements for wide-angle scanning of antenna beams have been met largely by using new light-weight, low-loss dielectrics to construct suitable lenses such as a spherical Luneberg lens. Using the geodesic analog of the Luneberg lens in one plane a scan of 40 beamwidths without aberrations has been obtained.

The very high gain required for antennas suitable for VHF ionospheric scatter propagation has been obtained through use of long horizontal rhombics and, more recently, by corner reflector antennas driven by collinear arrays of dipoles. The latter antennas have the important advantage for this application of much lower sidelobe level.

Traveling-wave antennas have received much attention in recent years because of their inherent adaptability to flush-mounted applications. Among the forms considered are corrugated surfaces and single or double dielectric-clad surfaces. The launching problem has been studied rather intensively and recent interest has been shown in the synthesis and scanning aspects for slow-wave structures. Progress has continued in the exploitation of fast-wave systems, with major advances centering on the launching problem and the polarization problem. Strip lines are taking an important place as feeding systems for traveling-wave antennas.

Recent years have been marked by considerable research activity on flush-mounted microwave antennas. Such antennas often take the form of slots or apertures in the metallic surfaces of aircraft. The radiation pattern of the slot radiator depends upon the shape of the metallic surface in which it is cut, and for complicated shapes resort is made to experiment. However, certain simple geometric shapes have been treated mathematically. These shapes include the circular cylinder, half plane, wedge, sphere, elliptic cylinder, oblate and prolate spheroid, and cones. Various combinations of these shapes on which work also has been done are the semicircular boss on a flat ground plane, the cylindrically tipped wedge, and the spherically tipped cone.

The problem of radiation from apertures in a metallic surface is closely related to the re-

rocal problem of calculating the currents ited on the surface by an incoming wave. a direct consequence of the reciprocity eorem there is an intimate connection be- een slot radiators, diffraction, and scattering. e result is that the extensive body of knowl- ge classed as diffraction and scattering must considered as an integral part of antenna eory. During the past three years there has en a very considerable effort in obtaining new act solutions, several new approximate solu- ns, and a better physical understanding of e mechanisms involved in scattering. All of e geometric shapes mentioned above have ceived attention, and much experimental ork has also been done. For the future it can expected that this intensive effort will con- nue with emphasis on the asymptotic ap- oaches of Kline and Keller, the method pro- sed by Logan, and Fock's method of obtain- g asymptotic results based on local analysis.

Communications

Preliminary Results of Measurements on oppler Shift of Satellite Emissions—P. R. rndt (p. 99)

Suppressed Sidelobe Antenna of 32 Ele- ents—Grote Reber (p. 101)

Measuring the Capacitance Per Unit Length Two Infinite Cones of Arbitrary Cross Sec- on—J. D. Dyson (p. 102)

The capacitance per unit length, and hence e characteristic impedance of two infinite nes, may be accurately measured by employ- g an extension of conventional guard tech- iques.

An appropriate gap is cut in one of the nes, converting the structure into a three- rminal capacitor. If the arm beyond the gap long enough, the desired field distribution ill be maintained past the gap and the capaci- ance per unit length of the isolated or guarded ection may be measured by a conventional w-frequency capacitance bridge.

The Exact Solution of the Field Intensities rom a Linear Radiating Source—Sheldon S. andler (p. 104)

Correction to "Determination of a Current Distribution over a Cone Surface Which Will Produce a Prescribed Radiation Pattern"—H. Unz (p. 104)

Fall Meeting of International Scientific Radio Union October 20–22, 1958, Pennsyl- ania State University (p. 105)

Abstracts of Papers from the Region Three Technical Meeting (p. 113)

Contributors (p. 115)

Audio

Vol. AU-6, No. 6, November–December, 1959

PGA News (p. 117)

An Improved Method for the Measurement of Nonlinear Audio Design—James A. Aagaard (p. 121)

The three most common methods of measuring nonlinear distortion in audio equipment are the harmonic method in which harmonics of a single sinusoidal input signal are measured; the SMPTE, or modulation, method in which the modulation of a high audio frequency by a low frequency is measured; and the CCIF, or difference frequency, method in which the beat note between two closely spaced frequencies is measured. These methods are discussed with particular reference to the behavior which may be expected in applications where the equipment under test includes pre-emphasis or de-emphasis networks, or where the distortion is symmetrical. It is shown that in a number of cases a more satisfactory method for use at the higher audio frequencies would be the measurement of the third-order component, rather than the second, produced in the CCIF method. A modification of the CCIF method using a sharp-cutoff low-pass filter is described which is capable of measuring both second- and third-order components. It is then shown that an instrument for this new method and the standard SMPTE method have many elements in common and that one instrument could be devised to make both forms of tests. The discussion is illustrated with the results of measurements made on a simulated distortion generator and on actual samples of audio equipment.

Modulation Noise in Two-Channel Disk Recordings—Daniel Cronin (p. 130)

Part of a modulation-noise problem encountered in two-channel stereo disk recording is shown to have its source not in the recording channel but in the geometry of the record-playback system. A relation is given to show the order of magnitude of this effect and some of the possible means of improvement are indicated.

Electromagnetic Efficiency of Heads in Magnetic Recording—Marvin Camras (p. 131)

A simple method is described for obtaining a figure of merit for response of magnetic recorder due to core losses and electrical characteristics.

Correspondence (p. 133)
Contributors (p. 133)
Annual Index 1958 (p. 136)

Electron Devices

Vol. ED-6, No. 1, January, 1959

On Calculating the Current Gain of Junction Transistors with Arbitrary Doping—H. L. Armstrong (p. 1)

In transistors made by such techniques as solid-state diffusion, the doping density is not constant in the emitter and base regions. Tanenbaum and Thomas have given an expression for the emitter efficiency of such a transistor. The present work extends that treatment, and gives an expression for the base transport factor. Thus the current gain can be calculated for any arbitrary distribution of doping density. It is found that, for constant doping densities, the results from this treatment reduce to those usually given (at least to the usual approximation), as they should.

Two cases of some practical interest are that in which the doping density varies monotonically from one side of the base to the other, as in the "drift transistor," and that in which the conductivity in the base region has a maximum somewhere within that region. Results for simplified forms of these two cases are presented in the form of graphs for convenience. Also, the result given by the present approximation for an exponential variation of doping density is compared with the exact calculation which can be made for this case.

Traveling-Wave Tube Efficiency Degradation Due to Power Absorbed in an Attenuator—C. K. Birdsall and C. C. Johnson (p. 6)

Traveling-wave tubes will oscillate if sufficient power is reflected at the input and the output of the tube; that is, if the loop gain is more than unity. As most tubes have large forward gain and matches (or loads) have some reflection, protection against oscillation is usually provided by return or backward loss. In most tubes the loss will be bilateral with the loss in the forward direction both absorbing power and reducing the rate of gain. The power absorbed subtracts from the power output and reduces the efficiency. This report presents the efficiency reduction due to power absorbed in the attenuator.

A Class of Waveguide-Coupled Slow-Wave Structures—J. Feinstein and R. J. Collier (p. 9)

The properties of an array of resonators coupled to a waveguide at their normally short-circuited ends are investigated. Methods of obtaining both normal and abnormal dispersion are indicated. Experimental measurements of dispersion and electronic impedance are compared with theoretical calculations. The effect of system parameters upon these characteristics is illustrated.

A Gun and Focusing System for Crossed-Field Traveling-Wave Tubes—O. L. Hoch and D. A. Watkins (p. 18)

A problem encountered in the design of crossed-field traveling-wave tubes (particularly M-type amplifiers or backward-wave oscillators) is a limitation on current due to the restricted cathode size in the usual gun systems. The scheme considered here incorporates into a crossed-field device the techniques commonly used in O-type devices for designing converging Pierce-type strip-beam guns, which can increase the effective cathode area by perhaps ten times. A design procedure is presented for getting a well-formed beam from such a gun, which is magnetically shielded, through a fringing crossed-field region into the uniform-field region of interaction.

An analysis of electron flow through the fringing-field region, including the effect of space charge, is presented. A trajectory equation, solved on a digital computer, yields trajectories for the beam and design curves for various values of the important parameters. The results show the scheme to be feasible.

Results are also presented from tests on an experimental, demountable tube used to test the focusing scheme. About 90 to 95 per cent of the current entering the crossed-field region could be focused to the collector, and the system behaved generally as the design predicted. An evaluation of the experimental data showed the scheme to be useful.

Germanium *P-N-P-N* Switches—I. A. Lesk (p. 28)

Transistors having bases from 90 to 220 mils wide, with the base contact placed near the collector, have shown marked current gain increases with current, peaking at values from 0.1 to 0.3. This effect is due to drift of the injected carriers in the base electric field set up by the emitter current. If the base contact is placed near the emitter, only very small values of α are obtained.

Germanium *p-n-p-n* diode and triode switches were made to operate primarily on the base field mechanism. They had one very wide (*p*-type) and one very narrow (*n*-type) base. The base contact for the switching triode may be placed either on the wide base but near the narrow base, or on the narrow base. In the latter case, much less current is required to switch the device from the high resistance to the high conductance region. When used as a transistor, the dc grounded base current gain of the *p-n-p-n* triode passes through unity at the turn-off current.

A *p-n-p-p+* structure, as a component part of the *p-n-p-n* triode with base contact on the wide base region, is shown to act as a switching diode in much the same way as the germanium *p-n-p-m* unit.

When the base contact is made to a narrow base bar-type transistor structure, overlap onto emitter and collector regions often results in a current gain that increases with current. This effect may also be utilized in making *p-n-p-n* switches.

Large-Signal Theory of UHF Power Triodes—A. D. Sutherland (p. 35)

Using a parallel plane model, an analysis is made of the UHF electronics of triodes operating under Class B conditions. The theory, in its most general form, includes the effects of space charge in both the grid-cathode region and the plate-grid region. However, most of the solutions contained in this paper were obtained by neglecting space-charge effects in the plate-grid region.

Over the wide range of UHF operating conditions investigated, it is found that the reduction of a triode's power output due to transit times is accounted for almost entirely by transit-time effects in the plate-grid region only. The principal effect of transit times in the grid-cathode region is to introduce a phase shift. This phase shift is quite important, however, for it contributes to a mechanism of regenerative feedback from output to input.

The reduction of transit-time effects for a fixed operating frequency is limited by four factors: temperature-limited saturation of the cathode emission current, reliable mechanical spacing considerations, electrode dissipation capabilities, and amplifier bandwidth requirements. It is shown that there exists an optimum plate-grid spacing which leads to the highest possible RF output power consistent with the required bandwidth of the amplifier.

The Design and Characteristics of a Megawatt Space-Harmonic Traveling-Wave Tube—M. Chodorow, E. J. Nalos, S. P. Otsuka, and R. H. Pantell (p. 48)

The tube described in this paper has a pulsed output power of one megawatt with 9.6 per cent bandwidth, where bandwidth is defined as the 3 db points for saturation power. Saturation gain is about 20 db, which is 6 db below the low-level gain. This is a space-harmonic *S*-band structure, designed to operate at a beam voltage of 10^5 volts and perveance of 2×10^{-6}. Field configurations for the lowest and next higher pass bands are drawn, based upon a field analysis and cold measurements. The impedance determined by perturbation measurements is compared to the impedance for the forced sinusoid inside a closed region, and it is found that the tube has about four times the minimum storage energy necessary to obtain the same bandwidth.

The Effect of a Space Charge on Bunching in a Two-Cavity Klystron—T. G. Mihran (p. 54)

The effect of space charge on bunching in a two-cavity klystron has been analyzed by many investigators, but no rigorous theory exists for large-signal bunching in a finite beam klystron.

The various available theories are reviewed and compared with new experimental data on the subject. It is found that there is 25 per cent loss of bunching efficiency at the drift length which is currently accepted as optimum, namely, 90 degrees of an effective plasma wavelength. It is shown that to reduce this loss to 5 per cent, drift length should vary from 65 degrees of an effective plasma wavelength at $\gamma b=0.5$ to 45 degrees at $\gamma b=2$.

Reasoning qualitatively from these measurements, the optimum drift lengths for a multicavity klystron are suggested to be:

$$90^\circ : 90^\circ : \cdots : 65^\circ\text{–}45^\circ : 30^\circ.$$

A Microwave Electron Velocity Spectrograph—P. B. Wilson and E. L. Ginzton (p. 64)

An electron-velocity spectrograph is described which can be used to measure the velocity distribution of electrons in a high-density electron beam modulated at microwave frequencies. The spectrograph employs a crossed electric and magnetic deflection system, has a resolution of less than one per cent in velocity, and a deflection region of very short length in order to minimize space-charge effects in the instrument.

Results of measurements made with the instrument on a Brillouin-focused, gap-modulated electron beam of microperveance one are given. Measurements made on the dc beam indicate that non-laminar flow effects are important for focusing fields of the order of and greater than the Brillouin focusing condition. Measurements of the RF velocity distribution at a point in the beam near the cavity gap have been found to agree with the predictions of kinematic bunching theory. A consequence of small-signal, space-charge-wave theory is that RF velocities should go to zero at a point in the beam a quarter of a reduced plasma wavelength from the cavity gap. This is confirmed by measurements made with this spectrograph.

Independent Space Variables for Small-Signal Electron Beam Analysis—D. L. Bobroff (p. 68)

This paper gives a systematic account of three systems of independent space variables useful in small signal electron beam analyses—the familiar Eulerian, polarization, and hose systems. The continuity, dynamical, and power exchange equations are given for each of the three sets of variables, and the relative merits of each system are discussed. The inapplicability of the Lagrangian system to small signal beam problems is also indicated.

Effect of Transient Voltages on Transistors—H. C. Lin and W. F. Jordan, Jr. (p. 79)

During transient conditions, the maximum junction temperature is dependent on the energy delivered to the transistor. For resistive and capacitive loads, transient energy is delivered during the turn-on period because the internal space-charge capacity induces a forward current in the base, which is amplified. For an inductive load, energy is delivered during the turn-off period because of the high induced voltage. Analytical expressions are derived to show the different transistor parameters and circuit constants which influence the magnitude of transient energy. By far the most important parameter is the sustain voltage. If this voltage is exceeded in the resistive or capacitive load condition, the initial base current is amplified by a greatly increased current gain causing excessive dissipation. If the voltage induced in the inductive load exceeds the sustain voltage, the negative resistance collector characteristic may cause highly dissipative oscillations which usually destroy the transistor. The transistor may be protected from destruction by preventing the base from biasing in the reverse direction so as to avoid negative resistance.

The Germanium Microwave Crystal Rectifier—Alan C. MacPherson (p. 83)

The characteristics of the commercial germanium microwave mixer crystal are reviewed. The conventional theory for these devices is discussed and criticized on the grounds that minority carrier storage phenomena are completely ignored. A model which includes this effect is proposed, and conversion loss calculations which include the effect of spreading resistance are presented for a highly idealized version of this model. Some possible explanations for the dc characteristics and for the success of the conventional fabrication techniques are offered.

Transient Analysis of Junction Transistors—W. F. Gariano (p. 90)

The transient behavior of the surface barrier and diffused junction transistors may be represented by an equivalent network consisting of two diodes and a nonlinear base resistance. Expressions for the nonlinear resistance, the base-emitter voltage time response, and the collector current time response are derived and tested experimentally. The equivalent circuit allows us to predict hole storage time for direct-coupled transistor logic (DCTL) circuits.

Electron-Beam Flow in Superimposed Periodic and Uniform Magnetic Fields—J. R. Anderson (p. 101)

Theoretical and experimental results are presented on the focusing of an electron beam by means of a magnet structure which produces, along the axis of the beam, a periodic magnetic field superimposed on a uniform field. The relation between space-charge and magnetic-field parameters for minimum ripple is derived. The flow in superimposed uniform and periodic magnetic fields is shown to be degraded from the flow of electrons in a magnetic field which has a sinusoidal variation along the axis. The results indicate the flow conditions to b[e] expected, where such combined fields are un[]avoidable. The focusing of electron beams i[n] this type of superimposed magnetic field and i[n] Brillouin flow are compared.

Strapped Bifilar Helices for High-Power Traveling-Wave Tubes—D. A. Watkins an[d] D. G. Dow (p. 106)

A method of increasing the peak-pulse[d] power output of broad-band traveling-wave tubes is described. The method involves the us[e] of a modified bifilar helix for the slow-wave structure. The modification employs 1) specia[l] straps or 2) mode-selective attenuation to pre[]vent backward-wave oscillation in the anti[]symmetric mode. This results in the possibilit[y] of using helices in the symmetric mode a[t] values of *ka* (circumference-to-free-space wave[]length ratio) as large as 0.6 at the highest amplification frequency. This in turn make[s] possible an increase in peak-pulsed beam powe[r] of a factor of approximately sixteen times tha[n] possible with a single helix. Both analytica[l] and experimental results regarding the be[]havior of the structures are presented showing the propagation characteristics. The experi[]mental results include cold measurements t[o] determine $\omega-\beta$ diagrams and measurements with an electron beam which yield experimenta[l] values of interaction impedance.

Dispenser Cathode Magnetrons—G. A. Espersen (p. 115)

The performance of a number of magnetrons using oxide nickel matrix L and impregnated types of cathodes is described. A discussion o[f] the cathode structure, evacuation, seasoning, emission factor, arcing and life is included.

Experimental Notes and Techniques

Method for Determining Specific Cooling Rates of Plate Materials in a Vacuum—C. W. Horsting, I. S. Solet, T. A. Sternberg, and P. Avakian (p. 119)

Contributors (p. 120)

Engineering Writing and Speech

Vol. EWS-2, No. 1, January, 1959

Industrial Electronics

Vol. IE-8, January, 1959

Message From the Publications Chairman (p. 1)

A Numerically Controlled Manufacturing System—F. E. Booth (p. 2)

The process of machining a finished part from a numerical description involves a number of data processing steps, as well as the actual metal cutting operations. The approach to ac

nplishing this purpose is generally organized two major groups of equipment. One group nsists of a general-purpose digital computer th associated input and output equipment. e second group of equipment consists of electronic machine control units, electrohydraulic vo drives, feedback units, and the machine ol itself. These equipment groups are deibed with reference to the Bendix system. so, operating experience in existing installans is discussed.

A High-Speed Low-Level Scanner—K. hslein (p. 17)

Pulse-Time Positioning Used for Safety ntrol of High-Speed Tester—M. E. Fitch . 25)

The operation and fail-safe features of the fety devices for a high speed bearing and shaft al testing machine, operating between 10,000 m and 110,000 rpm, are discussed in this per. The machine consists of an air turbine hich drives a load through a flexible quill aft. The safety devices interrupt the turbine r supply by a valve when either a predermined torque or a preset speed is exceeded. agnetic pickup signals generated by magtic pins attached to the rotating shaft operate e safety devices. An electrical shutoff signal developed during the first revolution of overrque or overspeed. These safety devices, inuding the valve, must be fail-safe and operate approximately 27 ms to prevent possible rbine disintegration due to centrifugal force the event of load removal by quill breakage.

A Positioning Servo for Automatic Machine ool Operation—F. H. London. (p. 31)

This paper describes a relay servo system sed in the positioning of machine tools. Its rimary application is in the manufacture of arts that have to fit accurately other parts on hich all machining has been completed. A near transducer uses the dimension of the achined part to obtain the reference signal, hile a matched linear transducer provides the rror signal for the servo system. A particular pplication of the servo system as used in a ully atomic railroad shop is described.

Automatic Packaging of Gelatin Capsules—. H. McMillan (p. 34)

Shipping cartons are automatically filled rom a 24 parallel channel feeder and photolectric detection system. A doubly preset ounter summarizes all 24 channels to first reluce speed and then stop the feed at a preset otal count. One conveyor system serves four uch feeder-counter combinations. The system as a counting capacity of approximately 00,000 capsules per minute.

A Nuclear Reactor Control System—D. J. Niehaus, R. R. Hoge, and A. B. Van Rennes p. 38)

Electronic equipment used with nuclear reactors performs the functions of operational control of the reactor, surveillance of the reactor power and period, and protection against langerous power excursions and periods. Transistorized circuits and duplicate channels of equipment provide the reliability which assures safe, continuous operation of the reactor.

Multichannel Swept Sonic Tester for Casting Quality Control—N. W. Schubring and J. E. Stevens (p. 46)

Automatic, production-type, sonic inspection based on the measurement of the fundamental vibrating frequency and/or decrement of a mechanically shocked casting has been in operation in our foundries for several years. The method has been suitable for the detection of cracks, voids, porosity, mottle, shrink, etc., on many of the parts having simple geometries. However, the more complex castings can vibrate in many basic modes and these modes upon shock excitation may interact indistinguishably. Swept- frequency, continuous-wave, forced vibration obviates mode interaction and permits casting evaluation by comparison of the sonic energy absorption spectra. Center frequencies, bandwidths, and amplitudes of certain of the absorption peaks usually establish overall quality.

The multichannel Swept Sonic Tester evaluates a given casting by determining the presence of a response within each of several preselected bands. These channels have very sharp limits which are established by active, tunable, twin-T networks. Each channel is self-contained permitting the system to be extended to accommodate any degree of casting complexity by adding channels in building block fashion. The spectrum sweep, having an adjustable sweep width, sweep rate, and starting frequency, is accomplished by heterodyning a crystal oscillator with a reactance-controlled frequency-swept oscillator and approximately follows the derived ideal frequency vs time characteristic for conserving inspection time.

Management Views Automation—I. Travis (p. 59)

Military Electronics

Vol. MIL-3, No. 1, January, 1959

Frontispiece—G. E. Valley (p. 1)

Guest Editorial—G. E. Valley (p. 2)

Systems Engineering and Weapon System Management—L. I. Davis (p. 4)

Systems Engineering for Usefulness and Reliability—W. C. Tinus and H. G. Och (p. 8)

With the increasing complexity and cost of weapon systems, it is becoming ever more important to provide a product that will be useful to the customer, that will provide reliable service, and that will have growth capabilities so that its useful life can be prolonged to meet the ever increasing enemy threat. The management of the research and development program for such large projects must provide detailed and careful planning and control in order to produce an integrated system on a minimum schedule.

System approach, now the byword of the electronic industry, means many things to many people. To the authors of this paper, it is the orderly arrangement of many details that are necessary to the sound planning of a large development effort.

Systems Engineering—R. H. Jewett and R. A. Montgomery (0. 12)

A description is given of practical systems engineering methods as applied to large military systems in an industrial environment. Particular emphasis is placed on a design approach which stresses minimum interconnections between subsystems and on system testing methods. Also discussed are system evaluation, management, and costs.

Weapons Systems Management—T. L. Phillips and I. A. Getting (p. 19)

Contributors (p. 23)

Ultrasonics Engineering

Vol. UE-7, February, 1959

PGUE Award (p. 1)

Biographical Notes on the Award Winners (p. 2)

An Ultrasonic Cleaner for Hypodermic Needles and Similar Small Bore Apparatus—F. Nesh and J. R. Andreotti (p. 3)

Most of the information dealing with ultrasonic cleaning devices is to be found in the advertising literature of the commercial houses manufacturing such equipment. G. G. Carr gives a rather comprehensive list of such concerns plus a discussion of the applications. Q. C. McKenna in his paper describes quite thoroughly the various techniques used in ultrasonic cleaning, with emphasis on its application to cleaning small parts. The ultrasonic equipment used at present for cleaning medical apparatus has not been effective in removing the coagulated blood from the inside of hypodermic needles. A design is described which has successfully removed the dried blood from hypodermic needles in a relatively short period of time. Preliminary experiments indicate that this system may also be sterilizing the needles at the same time as it is cleaning them.

Ultrasonic Atomization of Liquids—J. N. Antonevich (p. 6)

Observations of liquid films excited ultrasonically are described and discussed. These films atomized under ultrasonic excitation, and the rupture of capillary waves in the film is suggested as the prime cause of atomization. The atomization of a paint film by 20 kc vibrations was studied qualitatively. Highspeed motion pictures show sprays emanating from vibrating gas bubbles in the paint film. The particles produced were from 5μ to 500μ in diameter. Under the best atomizing conditions at 20 kc, the particle diameters were from 20μ to 70μ.

Theory of Magnetostrictive Delay Lines for Pulse and Continuous Wave Transmission—R. C. Williams (p. 16)

The theory presented is an analytic formulation of the characteristics of magnetostrictive delay lines. Assuming linear relationships between the elastic and magnetic variables involved, equations are derived that give analytic forms for the magnetostrictive and inverse magnetostrictive effect. The analytic expression for the magnetostrictive effect enables the strain pulse to be calculated in terms of the driving magnetic field. The expression for the inverse magnetostrictive effect gives the output magnetic flux density for an open-circuited receiver coil. The output voltage is then determined from this flux density.

Frequency response curves are obtained for two cases: 1) the ideal case, which refers to magnetic fields that terminate sharply at the extremes of the transmitting coil and receiving coils with perfect efficiencies; and 2) nonuniform and fringing magnetic fields at the transmitting coil and inefficient receiver coils. The response curves enable one to calculate the effective length of the coils needed to operate at the frequency which gives maximum gain. They also show that the nickel magnetostrictive delay line has an intrinsic conversion loss of 35 db. The voltage response is obtained by Fourier transform methods for input current step-functions for both of the above cases. These waveforms agree with experiment.

Biographical Notes on the Authors (p. 39)

Abstracts and References

Compiled by the Radio Research Organization of the Department of Scientific and Industrial Research, London, England, and Published by Arrangement with that Department and the *Electronic and Radio Engineer*, incorporating *Wireless Engineer*, London, England

NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger (†) must be regarded as provisional.

The Index to the Abstracts and References published in the PROC. IRE from February, 1958 through January, 1959 is published by the PROC. IRE, May, 1959, Part II. It is also published by *Electronic and Radio Engineer*, incorporating *Wireless Engineer*, and included in the March, 1959 issue of that journal. Included with the Index is a selected list of journals scanned for abstracting with publishers' addresses.

ACOUSTICS AND AUDIO FREQUENCIES

534:061.3 661
All-Union Acoustical Conference—V. A. Krasil'nikov. (*Akust. Z.*, vol. 4, pp. 105–106; January/March, 1958.) Report on a conference held in Moscow, June, 1957, at which 150 papers were read on propagation in inhomogeneous media, radiation and diffraction, waves of finite amplitude, ultrasonics, musical and physiological acoustics, and speech investigations.

534.2 662
Amplitude and Phase Fluctuations in a Spherical Wave—V. N. Karavaĭnikov. (*Akust. Z.*, vol. 3, pp. 165–176; April/June, 1957.) Mathematical analysis of fluctuations produced by inhomogeneities of the medium.

534.2 663
Correlation of Field Fluctuations—L. A. Chernov. (*Akust. Z.*, vol. 3, pp. 192–194; April/June, 1957.) Formulae are derived which establish a relation between the correlation function of the field fluctuation and the autocorrelation functions of the amplitude and phase fluctuations. See also 320 of 1957.

534.2–14 664
Diffraction and Radiation of Acoustic Waves in Liquids and Gases: Part 2—M. D. Khaskind. (*Akust. Z.*, vol. 4, pp. 92–99; January/March, 1958.) General expressions are derived for the average value of hydrodynamic forces and moments acting on a body in the presence of diffraction and radiation. Part 1: 3665 of 1958.

534.21 665
Waveguide Sound Propagation in One Type of Stratified Inhomogeneous Medium—Yu. L. Gazarian. (*Akust. Z.*, vol. 3, pp. 127–141; April/June, 1957.) An expression is derived for the field of a spherical harmonic point source in a medium in which the sound velocity varies according to Epstein's law (see 319 of 1957). A solution is given to waveguide-type propagation in an inhomogeneous half space with a totally reflecting boundary.

534.21 666
Acoustic Field in a Medium with a Homogeneous Surface Layer—A. N. Barkhatov. (*Akust. Z.*, vol. 4, pp. 13–18; January/March, 1958.) Experimental investigation of the propagation of sound through a space bounded by a homogeneous surface layer over a medium with a constant negative gradient of sound velocity.

534.21 667
Sound Amplitude Fluctuations in a Turbulent Medium—B. A. Suchkov. (*Akust. Z.*, vol. 4, pp. 85–91; January/March, 1958.) Report of an experimental investigation of the fluctuation of sound waves propagated through atmospheric layers near the ground.

534.21 668
On the Absorption of Sound Waves of Finite Amplitude—K. A. Naugol'nykh. (*Akust. Z.*, vol. 4, pp. 115–124; April/June, 1958.) Review of theory and comparison of results of calculations with experimental data, showing that waveform distortion, leads to a marked increase in absorption. In water, for example, at 100 kc, the absorption coefficient doubles for a pressure increase of the order of 0.01 atm.

534.21 669
Attenuation of a Sound Beam Traversing a Layer of Discontinuity in Sound Velocity—A. N. Barkhatov and I. I. Shmelev. (*Akust. Z.*, vol. 4, pp. 125–127; April/June, 1958.) A note on the experimental determination of the attenuation of a sound wave traversing a transition layer between two homogeneous media. The application of geometrical theory to the phenomena is considered.

534.21 670
Rayleigh-Type Waves on Cylindrical Surfaces—I. A. Victorov. (*Akust. Z.*, vol. 4, pp. 131–136; April/June, 1958.) Mathematical treatment of the propagation of Rayleigh-type elastic waves along a convex and a concave cylindrical surface.

534.21–8–14 671
Ultrasonic Absorption in Viscous Liquids—I. G. Mikhaĭlov. (*Akust. Z.*, vol. 3, pp. 177–182; April/June, 1957.) Ultrasonic absorption was measured in castor oil and other oils in the frequency range 0.26–30 mc. Results are tabulated.

534.232 672
On the Theory of Piezoelectric Transducers—K. V. Goncharov. (*Akust. Z.*, vol. 4, pp. 37–46; January/March, 1958.) Investigation of the frequency characteristics of X-cut quartz plates, steel, Al, fused quartz and Mg. The effect of an adhering layer is considered in relation to its thickness and acoustic properties.

534.232–8 673
Distributed Transducer—M. Greenspan and R. M. Wilmotte. (*J. Acoust. Soc. Am.*, vol. 30, pp. 528–532; June, 1958.) In an array of transducers separated by inactive material, the input voltages at successive inputs are delayed so that the speed of the electric wave travelling towards the load equals the speed of sound in the transducer material. High power and a wide frequency band can thus be obtained. See also 3012 of 1954 (Rabinow and Apstein).

534.232.001.4:534.522.1 674
Visualization of Mode Conversion of an Ultrasonic Beam in Fused Quartz—V. J. Hammond and R. Carter. (*Nature, London*, vol. 182, p. 790; September 20, 1958.) The process is useful in investigating methods of bonding transducers to fused quartz for use in delay lines.

534.26 675
Sound Scattering on Inhomogeneous Surfaces—Yu. P. Lysanov. (*Akust. Z.*, vol. 4, pp. 47–50; January/March, 1958.) Description of a mathematical method for the solution of a system of *n* equations determining the complex amplitude of waves scattered from a flat surface with periodically varying acoustic conductivity for the case of normal incidence.

534.26 676
Scattering of Sound by a Thin Rod of Finite Length—L. M. Lyamshev. (*Akust. Z.*, vol. 4, pp. 51–58; January/March, 1958.) Mathematical analysis of the scattering of a plane monochromatic sound wave by a thin finite rod of circular cross-section shows that vibrations of the rod can produce an angular variation of the scattering characteristics.

534.374–8 677
Ultrasonic Interference Filters with Variable Transmission Frequencies—B. D. Tartakovskiĭ. (*Akust. Z.*, vol. 3, pp. 183–191; April/June, 1957.) The general theory of ultrasonic interference filters of single- and multilayer type is developed. See 912 of 1953 (Curtis and Hadey).

534.522.1 678
Diffraction of Light by Large-Amplitude Ultrasonic Waves—I. G. Mikhaĭlov and V. A

utilov. (*Akust. Z.*, vol. 4, pp. 174–183; pril/June, 1958.) The light intensity distribution is calculated according to the diffraction maxima for different waveforms. The results of calculations taking account of phase modulation are in good agreement with experimental data. See also *Ibid.*, vol. 3, pp. 203–204; April/June, 1957.

4.6:621.385.83:537.228.1 679
An Electronic-Acoustical Converter—Yu. B. mennikov. (*Akust. Z.*, vol. 4, pp. 73–84; nuary/March, 1958.) Methods of rendering acoustical field visible are noted and a detailed treatment is given of the image-converter be in which a piezoelectric plate is scanned an electron beam. See 2184 of 1956 (Oshepkov *et al.*).

4.781 680
Masking of English Words by Prolonged owel Sounds—J. J. O'Neill and J. J. Dreher. (*J. Acoust. Soc. Am.*, vol. 30, pp. 539–543; une, 1958.) Results of tests are analysed.

4.79 681
Temporary Threshold Shift and Masking r Noise of Uniform Spectrum Level—J. D. Miller. (*J. Acoust. Soc. Am.*, vol. 30, pp. 517–22; June, 1958.) An empirical relation between masking and temporary threshold shift is xamined experimentally.

34.79 682
Temporary Threshold Elevation Produced y Continuous and 'Impulsive' Noises—V. Spieth and W. J. Trittipoe. (*J. Acoust. Soc. Am.*, vol. 30, pp. 523–527; June, 1958.)

34.833 683
Surface Absorption of Sound in Internally ined Ducts—R. Piazza. (*Alta Frequenza*, vol. 7, pp. 44–53; February, 1958.)

34.843 684
The Distribution of Normal Modes of Vibration in a Rectangular Room According to the Frequency Spectrum and Direction—Ma Da-Yu (D. Y. Maa). (*Akust. Z.*, vol. 4, pp. 168–73; April/June, 1958.) The angular distribution approaches a random one for an increase in he dimensions of the room, for a shift of the signal spectrum in the direction of higher frequencies, or for a widening of the frequency band.

534.844:534.6 685
Testing the 'Echo Parameter' Criterion in Room Acoustics by Means of Measurements of Syllable Intelligibility—H. Niese. (*Hochfreq. und Elektroak.*, vol. 66, pp. 70–83; November, 1957.) Measurements of the "echo parameter" (see 2658 of 1957) at various points in four different halls are compared with subjective intelligibility tests at the same points. A close correlation between the results of objective and subjective tests is found.

534.845 686
Acoustic Properties of Some Types of Sound-Absorbing Material—Z. N. Baranova and K. A. Velizhanina. (*Akust. Z.*, vol. 3, pp. 99–103; April/June, 1957.) The results of investigations are tabulated.

534.845:534.414 687
Slit Resonators as Low-Frequency Sound Absorbers—D. G. Ragavan. (*J. Inst. Telecommun. Engrs. India*, vol. 4, pp. 213–219; September, 1958.) Theoretical values of resonance frequency, bandwidth, and maximum absorption agree fairly well with values obtained in test chambers and studios.

534.846 688
Acoustical Design of the Alberta Jubilee Auditoria—T. D. Northwood and E. J. Stevens. (*J. Acoust. Soc. Am.*, vol. 30, pp. 507–516; June, 1958.) Impedance-tube and reverberation-chamber data were obtained for materials and components of the two auditoria. Measurements were made in the halls before their completion and concluded by a test concert.

534.851.089(083.74) 689
IRE Standards on Recording and Reproducing: Methods of Calibration of Mechanically Recorded Lateral Frequency Records, 1958—(PROC. IRE, vol. 46, pp. 1940–1946; December, 1958.) Standard 58 IRE 19. S1.

621.395.61 690
The Effect of Mechanical Vibrations on the Response of Various Types of Microphone—A. Chiesa. (*Alta Frequenza*, vol. 27, pp. 54–60; February, 1958.)

621.395.61 691
The Effect on a Receiving System of a Set of Independent Noise Sources Located on the Surface of a Sphere of Finite Radius—V. I. Klyachkin. (*Akust. Z.*, vol. 4, pp. 153–160; April/June, 1958.) The concepts of concentration coefficient and directivity characteristic as applied to a receiving system in the field of a large number of independent noise sources distributed over a continuous surface are discussed, and concentration coefficients for several types of receiving system are determined.

621.395.623.7:534.831 692
Methods of Generating High-Intensity Sound with Loud-speakers for Environmental Testing of Electronic Components Subjected to Jet and Missile Engine Noise—J. K. Hilliard and W. T. Fiala. (*J. Acoust. Soc. Am.*, vol. 30, pp. 533–538; June, 1958.)

621.395.623.8 693
On a Method for Increasing the Stability of Sound Amplification Systems—L. N. Mishin. (*Akust. Z.*, vol. 4, pp. 64–72; January/March, 1958.) Description of a system of acoustic feedback in which the phase of the feedback is varied continuously. A mathematical justification is given for the choice of a particular value of phase deviation. Experimental results are given.

621.395.625.3 694
Braking Action in Magnetic-Tape Recorders—G. Hartmann. (*Elektron. Rundschau*, vol. 12, pp. 45–49; February, 1958.) Consideration of the mechanics of the braking action shows that a constant braking moment is desirable. The shortcomings of practical methods are discussed, and a purely electrical braking system is suggested whose characteristics closely approach the ideal.

ANTENNAS AND TRANSMISSION LINES

621.372.2+621.396.11 695
Transmission and Reflection of Electromagnetic Waves in the Presence of Stratified Media—Wait. (See 939.)

621.373.2.09 696
Electric Waves on Delay Lines—F. Borgnis. (*Elektrotech. Z.*, vol. 79, pp. 383–385; June 1, 1958.) Propagation conditions for surface waves along plane slow-wave structures are examined.

621.372.221 697
A Novel Construction Concept for Linear Delay Lines—D. Elders. (IRE TRANS. ON COMPONENENT PARTS, vol. CP-4, pp. 24–28; March, 1957. Abstract, PROC. IRE, vol. 45, p. 1035; July, 1957.)

621.372.221 698
The Significance of Phase and Group Delay—F. Kirschstein and H. Krieger. (*Nachrichtentech. Z.*, vol. 11, pp. 57–60; February, 1958.) The importance of the phase velocity in assessing distortion in coil-loaded transmission lines is shown experimentally.

621.372.8 699
Critical Cross-Sections in Irregular Waveguides—B. Z. Katsenelenbaum. (*Dokl. Ak. Nauk SSSR*, vol. 123, pp. 53–56; November 1, 1958.) Mathematical analysis for the determination of the amplitude of any wave in a rectilinear irregular slanting waveguide with ideal walls.

621.372.8.002.2 700
Waveguide Manufacturing Techniques—T. Beardow. (*Brit. Commun. Electronics*, vol. 5, pp. 772–778; October, 1958.) Survey and comparison of techniques based on fabrication, casting, metal deposition, and printing methods.

621.372.821 701
The Characteristic Impedance and Phase Velocity of High-Q Triplate Line—K. Foster. (*J. Brit. IRE*, vol. 18, pp. 715–723; December, 1958.) An exact solution is obtained for the impedance in the absence of the dielectric support sheet and an expression for the phase velocity derived. Comparison with experimental results shows that the line parameters can be calculated with an error of about 1 per cent.

621.372.823 702
An Approximate Theory for Determining the Characteristic Impedances of Elliptic Waveguide—R. V. Harrowell. (*J. Electronics Control*, vol. 5, pp. 289–299; October, 1958.) Much similarity is found between the impedance behaviour of an elliptic waveguide sustaining an H_{C1} wave [see Chu, *J. Appl. Phys.*, vol. 9, pp. 583–591; September, 1938] and a rectangular guide sustaining an H_{10} wave. For a fixed broad dimension in each case the total axial wall current is not changed by reducing the length of the short dimension, and the impedance is modified only by the change in maximum voltage across the guide.

621.372.829 703
Helix Waveguide Theory and Application—H. G. Unger. (*Bell Syst. Tech. J.*, vol. 37, pp. 1599–1647; November, 1958.) Generalized telegraphist's equations are derived for the curved helix waveguide, and coefficients obtained for conversion from normal modes of this waveguide to normal modes of the metallic waveguide. A radial wave impedance at the helix interface is used to calculate the effect of composite jacket structures in three applications of circular-electric-wave transmission. The analysis is confirmed by measurement.

621.372.829 704
Attenuation of the TE_{01} Wave within the Curved Helix Waveguide—D. Marcuse. (*Bell Syst. Tech. J.*, vol. 37, pp. 1649–1662; November, 1958.) The helix waveguide has a coating of lossy dielectric and is shielded by a metallic pipe. A perturbation method is used to calculate the change in the field pattern of the TE_{01} mode caused by bending this waveguide; the additional field components produce an em field in the dielectric, which results in energy dissipation and attenuation. The attenuation can be reduced markedly by proper choice of dielectric thickness.

621.372.832.43 705
Modified Two-Hole Directional Coupler—W. G. Voss. (*Electronic Radio Eng.*, vol. 36, p. 28; January, 1959.) The modification described enables a two-hole, interference-type coupler to be tuned to give perfect directivity at any wavelength within a wide frequency range.

621.372.85 706
On the Theory of Anisotropic Obstacles in Waveguides—W. Hauser. (*Quart. J. Mech. Appl. Math.*, vol. 11, Part 4, pp. 427–437; November, 1958.) "Variational principles for the approximate computation of the elements of the scattering matrix for anisotropic obstacles in waveguides are presented."

621.372.852 707
Variational Principles or Guided Electromagnetic Waves in Anisotropic Materials—W. Hauser. (*Quart. Appl. Math.*, vol. 16, pp. 259–272; October, 1958.) Approximate expressions are given for propagation in a waveguide partially filled with a material with tensor electromagnetic properties. Variational principles are used to obtain a solution for a rectangular guide containing an infinitely long ferrite slab.

621.396.67:621.372.54 708
A Three-Bond Antenna Combining Network—Fife. (See 733.)

621.396.67.001.57 709
A Microwave Model Equipment for Use in the Study of the Directivity Characteristics of Short-Wave Antennas—D. W. Morris, E. W. Thurlow and W. N. Genna. (*P.O. Elec. Eng. J.*, vol. 51, pp. 126–131 and 173–179; July and October, 1958.) Report on investigations of the directivity characteristics of a rhombic antenna and a horizontal array of dipoles. The effect of nearby metallic structures on the performance of antennas was also studied.

621.396.677.4:621.396.11:551.510.535 710
A New Antenna to Eliminate Ground-Wave Interference in Inosopheric Sounding Experiments—R. S. Macmillan, W. V. T. Rusch, and R. M. Golden. (*J. Atmos. Terr. Phys.*, vol. 13, pp. 183–186; December, 1958.)

621.396.677.7:621.396.969.33 711
Slotted-Waveguide Array for Marine Radar—H. G. Byers and M. Katchky. (*Electronics*, vol. 31, pp. 94–96; December 5, 1958.) A rectangular waveguide with 126 inclined slots extends along the throat of a 30°-flare horn. A metal grating mounted across the horn aperture eliminates cross-polarisation radiation, and full-length chokes between waveguide and horn prevent back radiation. Advantages over a reflector-type antenna are noted.

621.396.677.8 712
Split Reflector for Microwave Antennas—R. L. Mattingley, B. McCabe, and M. J. Traube. (*Electronics*, vol. 31, pp. 86–88; December 19, 1958.) Impedance mismatch in pillbox (cheese) and other reflectors is reduced by dividing the reflector into two halves by a metal septum. Each half is fed by conjugate output ports of a short-slot hybrid coupler having suitable phase correction.

621.396.677.85 713
On the Axial Phase Anomaly for Microwave Lenses—G. W. Farnell. (*J. Opt. Soc. Amer.*, vol. 48, pp. 643–647; September, 1958.) Measurements of axial phase anomaly made on solid-dielectric lenses at 3 cm λ give results which generally agree with those calculated from scalar diffraction theory.

621.396.677.85 714
Experimental Investigation of a Homogeneous Dielectric Sphere as a Microwave Lens—R. N. Assaly. (*Can. J. Phys.*, vol. 36, pp. 1430–1435; October, 1958.) The emergent beam can be rotated through 360° by movement of the source alone. See also 996 of 1957 (Bekefi and Farnell).

AUTOMATIC COMPUTERS

681.142:061.4 715
Electronic Computer Exhibition—(*Wireless World*, vol. 65, pp. 17–21; January, 1959.) Short notes on special features of computers shown in London from November 28 to December 4, 1958.

681.142:061.4 716
Learning Machines—(*Wireless World*, vol. 65, pp. 8–9; January, 1959.) A conditional-probability computer and character-recognition machines, exhibited at the symposium on "The Mechanization of Thought Processes" held at the National Physical Laboratory from November 24–27, 1958, are discussed.

681.142:537.227 717
Ferroelectric Storage Devices—S. Morleigh. (*Electronic Eng.*, vol. 30, pp. 678–684; December, 1958.) The switching characteristics of ferroelectric materials, as required for use in digital storage systems, are examined. The properties of single-crystal $BaTiO_3$ capacitors are investigated experimentally.

681.142:621.318.57 718
A Track Switching System for a Magnetic Drum Memory—D. D. Majunder. (*Electronic Eng.*, vol. 30, pp. 702–705; December, 1958.)

CIRCUITS AND CIRCUIT ELEMENTS

621.3.077.6:621–526 719
Phase-Selective Gate Rejects Quadrature—B. Fennick. (*Electronics*, vol. 31, pp. 92–93; December 19, 1958.) A phase-reference voltage controls two unmatched diodes which conduct only when the in-phase component is at maximum and the quadrature component at minimum. A circuit designed for a 400 cps servo amplifier is described.

621.314.22 720
Transformer Design for Zero Phase Shift—N. R. Grossner. (IRE TRANS. ON COMPONENT PARTS, vol. CP-4, pp. 82–85; September, 1957. Abstract, PROC. IRE, vol. 46, p. 384; January, 1958.)

621.316.825 721
Bounds for Thermistor Compensation of Resistance and Conductance—A. B. Soble. (IRE TRANS. ON COMPONENT PARTS, vol. CP-4, pp. 96–101; September, 1957. Abstract, PROC. IRE, vol. 46, p. 384; January, 1958.)

621.318.57:621.314.7 722
An Introduction to the Use of Transistors in Inductive Circuits—A. F. Newell. (*Mullard Tech. Commun.*, vol. 4, pp. 157–160; November, 1958.) The mechanism of a delayed switch-off effect, which is due to avalanche multiplication, is described, and a method of avoiding it in relay switching circuits is given.

621.318.57:621.314.7 723
Transistor Switch Design—A. Gill. (*Electronics*, vol. 31, p. 97; December 5, 1958.) Switching parameters are tabulated for eight types of transistor.

621.318.57:621.387 724
An Experimental Gas-Diode Switch—A. D. White. (*Bell Lab. Rec.*, vol. 36, pp. 446–449; December, 1958.) The ac impedance is a stable negative resistance of about 225 Ω for frequencies up to 30 kc.

621.319.4 725
Dielectric Films in Aluminium and Tantalum Electrolytic and Solid Tantalum Capacitors—J. Burnham. (IRE TRANS. ON COMPONENT PARTS, vol. CP-4, pp. 73–82; September, 1957. Abstract, PROC. IRE, vol. 46, p. 384; January, 1958.)

621.319.4:621.318.134 726
Ferrite-Cored Capacitors—R. Davidson. (*Research, London*, vol. 11, pp. 367–370; September, 1958.) Details are given of British capacitors with ferrite loading for improving their attenuation characteristics.

621.372.029.3(083.7) 727
IRE Standards on Audio Techniques: Definitions of Terms, 1958—(PROC. IRE, vol. 46, pp. 1928–1934; December, 1958.) Standard 58 IRE 3. S1.

621.372.2 728
A Review on the Analysis of Transients in Electrical Circuits Using the Laplacian Transformation—P. R. Rao. (*J. Inst. Telecommun. Engrs. India*, vol. 4, pp. 209–212; September, 1958.)

621.372.44:621.316.8 729
General Power Relationships for Positive and Negative Nonlinear Resistive Elements—R. H. Pantell. (PROC. IRE, vol. 46, pp. 1910–1913; December, 1958.) The method developed by Manley and Rowe (2988 of 1956) for the treatment of reactive elements is extended to resistors. Relations are derived which yield modulation efficiency, efficiency of harmonic generation and stability criteria.

621.372.5 730
How Quickly does a Twin-T respond?—M. Price. (*Can. Electronics Eng.*, vol 2, pp. 40–41; September, 1958.) The response to three different input waveforms is examined using the Laplace transformation. In all cases the output transient is negligible after $1\frac{1}{2}$ cycles of the resonance frequency of the network.

621.372.54 731
Filter Attenuation Characteristics—M. D. Johnson and D. A. G. Tait. (*Electronic Eng.*, vol. 30, pp. 710–711; December, 1958.) The use of hyperbolic functions is avoided, and formulae are derived which permit slide-rule computation.

621.372.54:621.376.3 732
The Magnitude of the Permissible Circuit Impedances of the Filters in I.F. Amplifiers for F.M. Systems—E. G. Woschni. (*Hochfreq. und Electroak.*, vol. 66, pp. 63–67; November, 1957.) Application of results obtained in 2672 of 1958.

621.372.54:621.396.67 733
A Three-Band Antenna Combining Network—S. L. Fife. (*Electronic Eng.*, vol. 30, pp. 720–722; December, 1958.) Three prototype filter sections of low-pass, band-pass and high-pass characteristics respectively for bands 1, 2, and 3 are combined in a masthead unit to provide a common output for the three antenna inputs.

621.372.543 734
Normalized Input-Admittance Curves of Two-Stage Band Filters and the Smallest Mismatch Circles for a Given Frequency Range—H. Hein. (*Nachrichtentech. Z.*, vol. 11, pp. 85–91; February, 1958.)

621.372.543 735
Band Filters with Electronic Bandwidth Control—C. Kurth. (*Elektron. Rundschau*, vol. 12, pp. 39–44; February, 1958.) The design of a two-stage filter with amplification in the feedback line is considered.

621.372.543 736
Null-Point Band Filters and Their Theoretical Treatment—E. Trzeba. (*Hochfreq. und Elektroak.*, vol. 66, pp. 90–94 and 95–107; November, 1957, and January, 1958.)

621.372.553:621.397.61 737
The Frequency and Time Characteristics of

All-Pass Filters for Delay Equalization in Television Transmission—H. Dobesch. (*Hochfreq., und Elektroak.*, vol. 66, pp. 67–70; November, 1957.) The basic circuits given by Bünemann (2240 of 1956) are considered.

621.372.632:621.3.072.6 738

Some Properties of a Frequency Stabilizing Circuit—L. L. Campbell. (IRE TRANS. ON COMMUNICATIONS SYSTEMS, vol. CS-5, pp. 10–12; September, 1957. Abstract, PROC. IRE, vol. 45, p. 1760; December, 1957.)

621.373 739

The Wave-Mechanical Damped Harmonic Oscillator—K. W. H. Stevens. (*Proc. Phys. Soc.*, vol. 72, pp. 1027–1036; December 1, 1958.) A wave-mechanics treatment is given of damped harmonic motion for the charge in a tuned circuit and the em field in a resonant cavity. The classical frequency appears in the time dependence of the eigenfunctions, and the classical damping as a decay in the eigenvalues.

621.373.14:621.396.96 740

High-Q Echo Boxes—A. Cunliffe and R. N. Gould. (*Electronic Radio Eng.*, vol. 36, pp. 29–33; January, 1959.) An investigation of the occurrence of unwanted modes in a tunable Ho cylindrical cavity. Suggestions for the suppression of these modes are given.

621.373.421.13 741

Generations of Oscillations with Equally Spaced Frequencies in a Given Band—D. Makow. (IRE TRANS. ON COMMUNICATIONS SYSTEMS, vol. CS-5, pp. 13–20; September, 1957. Abstract, PROC. IRE, vol. 45, p. 1760; December, 1957.)

621.373.43:621.396.96 742

The Application of Pulse-Forming Networks—A. Graydon. (IRE TRANS. ON COMPONENT PARTS, vol. CP-4, pp. 7–13; March, 1957. Abstract, PROC. IRE, vol. 45, p. 1035; July, 1957.)

621.373.431 743

Miller Sweep Circuit—C. S. Speight. (*Wireless World*, vol. 65, pp. 34–36; January, 1959.) Circuits are described in which both Miller integrator and Puckle flyback circuits are combined to produce a very linear time base.

621.373.431:621.314.7 744

Bistable Circuits using Unijunction Transistors—T. P. Sylvan. (*Electronics*, vol. 31, pp. 89–91; December 19, 1958.) The design and operation of various circuits are explained. Use of the negative-resistance region as one stable state decreases power requirements and increases switching speed. The application to ring-counter circuits is shown.

621.373.52 745

Transistor 20-kc/s Oscillator with 50mW Output—J. F. Berry and L. E. Jansson. (*Mullard Tech. Commun.*, vol. 4, pp. 122–127; November, 1958.) "A design procedure is described for transistor oscillators employing external feedback and an *LC* resonator. An example is given which delivers an output of 50m at 20kc, and which operates from a 9-volt battery. It is suitable for use as a bias oscillator for dictation machines."

621.373.52 746

Graphical Designing of Transistor Oscillators—W. R. McSpadden and E. Eberhard. (*Electronics*, vol. 31, pp. 90–93; December 5, 1958.) A method is described which is simple but yields results accurate enough for most engineering design calculations. A design example is given for a crystal-controlled oscillator to operate at 1 mc.

621.375.018.756 747

Summary of the Theory of Wide-Band Distributed Amplifiers Suitable for the Amplification of Very Short Pulses—D. Dosse. (*Nachrichtentech. Z.*, vol. 11, pp. 61–68; February, 1958.)

621.375.024:621–52 748

D.C. Amplifiers for Control Systems—L. S. Klivans. (*Electronics*, vol. 31, pp. 96, 98; November 21, 1958.)

621.375.121.2:621.396.621.22 749

An Electronic Multicoupler and Antenna Amplifier for the V.H.F. Range—K. Fischer. (IRE TRANS. ON COMMUNICATIONS SYSTEMS, vol. CS-5, pp. 43–48; December, 1957. Abstract, PROC. IRE, vol. 46, p. 515; February, 1958.)

621.375.2.024 750

A Triode-Connected Pentode with Stabilized Anode Current—B. C. Cox. (*J. Sci. Instr.*, vol. 35, pp. 471–472; December, 1958.) Drift in dc coupled circuits caused by heater supply voltage variation is minimized by the use of a triode-connected pentode stabilized by variation of grid current with emission.

621.375.2.132.3 751

The Influence of the Output Time-Constant of a Cathode Follower—C. Edwards. (*Electronic Eng.*, vol. 30, pp. 712–714; December, 1958.) A nomogram is derived for determining the maximum pulse amplitude, as a function of rise time and output time constant, that may be applied without causing positive grid current.

621.375.23:621.317.089.2 752

A Low-Capacitance Input Circuit—J. C. S. Richards. (*Electronic Eng.*, vol. 30, pp. 706–708; December, 1958.) Methods of reducing the effect of coaxial-cable capacitance are discussed. A probe containing only passive elements is used with a single triode-pentode valve feedback amplifier to give an input capacitance of 5 pf and a gain of unity over a bandwidth of 2 Mc.

621.375.4:621.318.57 753

Properties of Hook Transistors in Switching and Amplifying Circuits—L. M. Vallese. (*J. Brit. IRE*, vol. 18, pp. 725–732; December, 1958.) An analysis is given of the most significant properties of hook and *p-n-p-n* transistors in common-base, common-emitter and common-collector configurations.

621.375.4.029.3 754

Stagger-Tuned Transistor Video Amplifiers—V. H. Grinich. (IRE TRANS. ON BROADCAST AND TELEVISION RECEIVERS, vol. BTR-2, pp. 53–56; October, 1956. Abstract, PROC. IRE, vol. 45, p. 253; February, 1957.) See also 69 of January.

621.375.4.078 755

Transistor Circuits Based on the Half-Supply-Voltage Principle—B. G. Dammers, A. G. W. Uitjens, and W. Ebbinge. (*Electronic Applic.*, vol. 18, pp. 85–98; August, 1958.) Analysis shows that temperature stabilization of a transistor can be effected by a preceding dc coupled amplifying stage stabilized on the half-supply-voltage principle. Practical circuits are described for AF amplifiers and stabilized supply units. See also 71 of January (Dammers et al.).

621.375.9:538.569.4.029.64 756

Analysis of the Emissive Phase of a Pulsed Maser—H. H. Theissing, F. A. Dieter, and P. J. Caplan. (*J. Appl. Phys.*, vol. 29, pp. 1673–1678; December, 1958.) A discussion of the emission from a matched cavity at paramagnetic resonance. Results are given for transverse moment, output field, output power, and power gain as a function of time.

621.375.4+621.373.52 757

Transistor Circuit Engineering [Book Review]—R. F. Shea (ed.) Publishers: John Wiley, New York, and Chapman and Hall, London, 1957, 468 pp., 95 s. (*Nature, London*, vol. 182, pp. 756–757; September 20, 1958.)

GENERAL PHYSICS

537+538]:061.3 758

Report of the Meeting of the Swiss Physical Society—(*Helv. Phys. Acta*, vol. 30, pp. 457–494; November 20, 1957.) The text is given of papers read at a meeting held at Neuchâtel, September 1957, including the following:

a) **Electrical Properties of Silver Selenide Ag_2Se**—G. Busch and P. Junod (p. 470, in French).

b) **Oscillatory Magnetic Resistance Variation of *n*-Type InSb at Low Temperatures and High Field Strengths**—G. Busch, R. Kern and B. Lüthi (pp. 471–472, in German.)

c) **The Field Parameters of Galvano- and Thermomagnetic Effects in Ferromagnets**—G. Busch, F. Hulliger, and R. Jaggi (pp. 472–474, in German).

d) **Hall and Righi-Leduc Effects in Ferromagnetics**—D. Rivier (pp. 474–478, in French).

e) **Effect of a Cubic Electric Field on the Fundamental Level of the Gd^{+++} Ion**—R. Lacroix (pp. 478–480, in French).

f) **Hyperfine Splitting in the Paramagnetic Resonance of Pr^{3+} in Ceramic $LaAlO_3$**—H. Gränicher, K. Hübner, and K. A. Müller (pp. 480–483, in German).

g) **Improvements in an NH_3 Maser**—J. Bonanomi, J. De Prins, J. Herrmann, and P. Kartaschoff (pp. 492–494, in German).

537.122 759

Magnetic Susceptibility of an Electron Gas at High Density—K. A. Brueckner and K. Sawada. (*Phys. Rev.*, vol. 112, pp. 328–329; October 15, 1958.)

537.222 760

The Electrostatic Interaction of Two Arbitrary Charge Distributions—M. E. Rose. (*J. Math. Phys.*, vol. 37, pp. 215–222; October, 1958.)

537.226.1 761

Molecular Theory of the Dielectric Constant—L. Jansen. (*Phys. Rev.*, vol. 112, pp. 434–444; October 15, 1958.) A theory of the static dielectric constant is developed on a quantum-mechanical basis. By introducing the concept of "local field" a molecular version of the general theory is obtained. It is shown that such a molecular theory is fundamentally ineffective in accounting for the observed results within the experimental accuracy.

537.311.31:530.145 762

Quantum Theory of the High-Frequency Conductivity of Metals—M. Ya. Axbel'. (*Zh. Eksp. Teor. Fiz.*, vol. 34, pp. 969–983; April, 1958.) Development of the theory of conductivity of a metal in a high-frequency em field and a constant magnetic field. The amplitude of the quantum oscillations in the high-frequency case is usually larger than in the static case.

537.311.62 763

Quantum Oscillations of the High-Frequency Surface Impedance—M. Ya. Azbel'. (*Zh. Eksp. Teor. Fiz.*, vol. 34, pp. 1158–1168; May, 1958.) A quantum-mechanical formula is derived and cases involving constant magnetic fields both parallel and inclined to the surface are considered. From an experimental study of surface impedance in a strong magnetic field the shape of the Fermi surface and the velocity of the electrons in it can be determined.

537.533:621.385.029.6 764

Solution to the Equations of Space-Charge Flow by the Method of the Separation of Variables—P. T. Kirstein and G. S. Kino. (*J. Appl. Phys.*, vol. 29, pp. 1758–1767; December, 1958.) The steady-state behaviour of electron beams with high space-charge densities is analysed. The equations for irrotational, electrostatic laminar space-charge flow are set up in terms of the action function; these equations are reduced by separation of variables in cylindrical polar coordinates. The method may be extended to include the effect of magnetic fields.

537.56 765

Energy Spectrum of Plasma Electrons—G. Medicus. (*Elektrotech. Z.*, vol. 79, pp. 373–374; June 1, 1958.) Brief report on results obtained by the graphical method described earlier (731 of 1957).

537.56 766

Thermal Conductivity of an Electron Gas in a Gaseous Plasma—T. Sekiguchi and R. C. Herndon. (*Phys. Rev.*, vol. 112, pp. 1–10; October 1, 1958.) Experimental techniques used with Ne and He plasmas are described. Conductivity values obtained by two separate methods are in agreement and consistent with theory.

537.56:538.56 767

Kinetic Theory of Magnetohydrodynamic Waves—K. N. Stepanov. (*Zh. Eksp. Teor. Fiz.*, vol. 34, pp. 1292–1301; May, 1958.) Analysis of the propagation of magnetohydrodynamic waves in an ionized gas when the wave frequency is much greater than the frequency of "short-range" collisions. See also 2418 of 1957.

537.562 768

Microwave Method for Measuring the Probability of Elastic Collision of Electrons in a Gas—J. L. Hirshfield and S. C. Brown. (*J. Appl. Phys.*, vol. 29, pp. 1749–1752; December, 1958.) A plasma in a dc magnetic field has a transverse conductivity component whose reactive part depends on the magnetic field. By measuring the magnetic field necessary to make the reactive part zero the probability of the elastic collision of electrons in he is obtained.

537.562:538.56 769

Scattering of Microwave Radiation by a Plasma Column—F. I. Boley. (*Nature, London*, vol. 182, pp. 790–791; September 20, 1958.) Experiments conducted at 10 cm λ to determine the angular distribution of the e.m. radiation scattered by the positive column of a mercury discharge under resonant conditions are described. Results support the theory of Mackinson and Slade (3532 of 1954). See also 418 of 1958 (Dattner).

538.24:621.372.413 770

Microwave Faraday Rotation: Design and Analysis of a Bimodal Cavity—A. M. Portis and D. Teaney. (*J. Appl. Phys.*, vol. 29, pp. 1692–1698; December, 1958.) An equivalent circuit is developed for the cavity, and the coupling between degenerate modes is expressed in terms of the susceptibility tensor of the material producing the rotation. The results are compared with experimental data.

538.566 771

Propagation of Plane Electromagnetic Waves in Inhomogeneous Media—H. Osterberg. (*J. Opt. Soc. Amer.*, vol. 48, pp. 513–521; August, 1958.) The laws of propagation along the z direction are derived for infinite inhomogeneous media in which dielectric constant and electrical conductivity are functions of z and the magnetic permeability is constant. Homogeneous media are treated as special cases.

538.566 772

Electromagnetic Scattering by Thin Conducting Plates at Glancing Incidence—J. S. Hey and T. B. A. Senior. (*Proc. Phys. Soc.*, vol. 72, pp. 981–995; December 1, 1958.) A large signal is scattered back from a thin plate illuminated edge-on by a field whose magnetic vector is normal to the plate. Experimental measurements show that the currents in the plate are predominant near the edges; their relation to theoretical results is discussed.

538.566:516.6 773

A Helical Coordinate System and Its Applications in Electromagnetic Theory—R. A. Waldron. (*Quart. J. Mech. Appl. Math.*, vol. 11, Part 4, pp. 438–461; November, 1958.) The system described enables problems involving helical symmetry to be solved exactly.

538.566:535.42 774

The Edge Condition in Diffraction Problems—P. Poincelot. (*Compt. Rend. Acad. Sci., Paris*, vol. 246, pp. 3324–3325; June 16, 1958.) The diffraction of an em wave by a perfectly conducting solid is considered.

538.566:535.43]+534.26 775

Theory of Wave Scattering on Periodically Uneven Surfaces—Yu. P. Lysanov. (*Akust. Z.*, vol. 4, pp. 3–12; January/March, 1958.) Description of six approximate mathematical methods for calculating the scattering of sound or em waves over the sea or uneven ground. 79 references.

538.566.2:538.22 776

Magnetic Double Refraction of Microwaves in Paramagnetics—F. S. Imamutdinov, N. N. Neprimerov, and L. Ya. Shekun. (*Zh. Eksp. Teor. Fiz.*, vol. 34, pp. 1019–1021; April, 1958.) Investigation at 9375 mc of the rotation of the plane of polarization of an H_{11} wave in a circular waveguide containing a paramagnetic salt, and its dependence on the intensity of the static magnetic field perpendicular to the direction of wave propagation.

538.569.4:538.221 777

Spin-Wave Analysis of Ferromagnetic Resonance in Polycrystalline Ferrites—E. Schlömann. (*J. Phys. Chem. Solids*, vol. 6, pp. 242–256; August, 1958.) Dipolar interaction is taken into account by means of the spin-wave formalism. Crystalline anisotropy and the polycrystalline nature of the material cause the homogeneous mode of precession to interact with spin waves whose wavelength is of the order of, or larger than, the average linear grain size. The theory predicts a very strong frequency and shape dependence of the line width when the homogeneous mode is approximately degenerate with long-wavelength spin waves propagating normal to the dc field.

538.569.4:538.222 778

Paramagnetic Resonance—J. S. van Wieringen. (*Philips Tech. Rev.*, vol. 19, pp. 301–313; May 31, 1958.) The quantum-mechanical theory of the phenomenon is discussed with reference to results of experimental investigations.

538.569.4.029.6:539.2 779

Technical Applications of Microwave Physics—D. J. E. Ingram. (*Research, London*, vol. 11, pp. 401–407; October, 1958.) Particular reference is made to electron-resonance techniques.

538.569.4.029.64 780

Paramagnetic Resonance Spectrum of Gadolinium in Hydrated Lanthanum Trichloride—M. Weger and W. Low. (*Phys. Rev.*, vol. 111, pp. 1526–1528; September 15, 1958.) "The paramagnetic resonance spectrum of Gd^{3+} in $LaCl_3 \cdot 7H_2O$ was measured and found to agree quite well with a spin Hamiltonian with dominant coefficients $b_2^0 = \pm 0.0131$ cm^{-1}, $b_2^2 = \mp 0.0075$ cm^{-1} at room temperature, and $b_2^0 = \pm 0.0099$ cm^{-1}, $b_2^2 = \mp 0.0115$ cm^{-1} at liquid air temperature."

539.2 781

Theory of Plasma Resonance in Solids—P. A. Wolff. (*Phys. Rev.*, vol. 112, pp. 66–69; October 1, 1958.) The modes of a confined plasma are studied for simple geometries. Modes are closely spaced in frequency and unresolvable unless the sample size is comparable to the Debye length. Observation in small samples is made difficult by line broadening due to surface scattering but might be possible in a suitably designed experiment.

539.2:548.0 782

Fourier Coefficients of Crystal Potentials—J. Callaway and M. L. Glasser. (*Phys. Rev.*, vol. 112, pp. 73–77; October 1, 1958.) A method is developed for the calculation of the Fourier coefficients of the electrostatic potential of a given distribution of valence electrons in a solid, taking full account of the nonspherical character of the atomic polyhedron.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.14:538.69:523.165 783

Interplanetary Magnetic Field and its Control of Cosmic-Ray Variations—J. H. Piddington. (*Phys. Rev.*, vol. 112, pp. 589–596; October 15, 1958.) A model interplanetary magnetic field is described which may explain some features of solar cosmic-ray increases and also fluctuations in the primary radiation. An attempt is made to show how localized solar fields may create the field, which should be largely radial in form.

523.15 784

A Theorem on Force-Free Magnetic Fields—L. Woltjer. (*Proc. Nat. Acad. Sci.*, vol. 44, pp. 489–491; June 15, 1958.) A variational principle is proved which provides a more direct and satisfactory approach than that given in 3788 of 1958 (Chandrasekhar and Woltjer).

523.164.32 785

Nonuniformity in the Brightness of the Sun's Disk Sunspot Minimum—J. C. Bhattacharyya. (*J. Atmos. Terr. Phys.*, vol. 13, pp. 43–44; December, 1958.) A comparison is made between the nonuniformity, as derived from ionospheric measurements during eclipses, observed in 1944 and 1954. The results suggest that there are significant differences for the two epochs.

523.164.32:523.75 786

Observation of a Solar Flare at 4.3-mm Wavelength—R. J. Coates. (*Nature, London*, vol. 182, p. 861; September 27, 1958.) Scans made across the sun on September 25–27, 1957, with a radio telescope ((1126*i* of 1958) are compared with corresponding scans for a quiet sun.

523.164.32:551.510.535 787

Solar Radiation on Decimetre Waves as an Index for Ionospheric Studies—M. R. Kundu and J. F. Denisse. (*J. Atmos. Terr. Phys.*, vol. 13, pp. 176–178; December, 1958.) The solar noise radiation flux at 10.7 cm λ is compared with other indexes of solar activity. It seems to be as good as any other index for ionospheric studies on a monthly time scale and better for shorter time intervals.

523.164.4 788

The Trapping of Cosmic Radio Waves beneath the Ionosphere—G. R. Ellis. (*J. Atmos. Terr. Phys.*, vol. 13, pp. 61–71; December, 1958.) When there is a horizontal gradient in

the critical frequency of a layer, incoming extraterrestrial radiation may be trapped between the layer and the ground and propagated over large distances. This would account for the reception of cosmic noise at frequencies lower than the local critical frequency.

523.164.4:523.74 789

Sudden Cosmic Noise Absorption Associated with the Solar Event of 23 March 1958—K. A. Sarada. (*J. Atmos. Terr. Phys.*, vol. 13, pp. 192–194; December, 1958.)

550.38 790

The Relationships between the Secular Change and the Non-dipole Fields—K. Whitham. (*Can. J. Phys.*, vol. 36, pp. 1372–1396; October, 1958.) The drift and decay contributions to the secular variation have been estimated from isomagnetic and isoporic charts for 1955 in Canada. The westward drift in recent years was found to be significantly smaller than the world-wide average. Relations are obtained between the Gaussian coefficients in the spherical harmonic analyses of the earth's main field and the secular variation. It is shown that one half of this variation is produced by westward drift and that the decay terms are unimportant.

550.389.2:551.510.535 791

Electron Density Profiles in the Ionosphere during the I.G.Y.—(*J. Atmos. Terr. Phys.*, vol. 13, pp. 195–197; December, 1958.) See 438 of February (Smith-Rose).

550.389.2:629.19 792

Use of Artificial Satellites to Explore the Earth's Gravitational Field: Results for Sputnik II (1957β)—R. H. Merson and D. G. King-Hele. (*Nature, London*, vol. 182, pp. 640–641; September 6, 1958.)

550.389.2:629.19 793

Seasonal Illumination of a Circumpolar Earth Satellite at its Extreme-Latitude Orbit Point—W. N. Abbott. (*Nature, London*, vol. 182, pp. 651–652; September 6, 1958.)

550.389.2:629.19 794

Polyhedral Satellite for More Accurate Measurement of Orbit Data of Earth Satellites—D. R. Herriott. (*J. Opt. Soc. Amer.*, vol. 48, pp. 667–668; September, 1958.) A 270-face polyhedron would reflect to an observer a light pulse of intensity more than 3000 times that from a sphere.

550.389.2:629.19 795

Rotation of Artificial Earth Satellites—R. N. Bracewell and O. K. Garriot. (*Nature, London*, vol. 182, pp. 760–762; September 20, 1958.) Radio signals received from Sputnik I were subject to deep regular fading with a semiperiod of about 4s, probably due to free motion of the satellite about its centroid. These fluctuations were less noticeable in the field strength record of Sputnik III. The advantages of a disc-shaped satellite and methods of eliminating rotational effects are discussed.

550.389.2:629.19 796

The Faraday—Rotation Rate of a Satellite Radio Signal—S. A. Bowhill. (*J. Atmos. Terr. Phys.*, vol. 13, pp. 175–176; December, 1958.)

550.389.2:629.19 797

An Irregularity in the Atmospheric Drag Effects on Sputniks II and III (Satellites 1957β, 1958δ' and 1958δ_2)—D. G. King-Hele and D. M. C. Walker. (*Nature, London*, vol. 182, pp. 860–861; September 27, 1958.)

550.389.2:629.19 798

Satellite Tracking by H.F. Direction Finder—J. L. Wolfe. (*J. Atmos. Terr. Phys.*, vol. 13, pp. 155–164; December, 1958.) A twin-channel cathode-ray DF with an Adcock aerial was used on the 20 mc signals from satellites 1957α and β. The results obtained and their accuracy are discussed.

550.389.2:629.19:551.510.535 799

Comparison of Phase Difference and Doppler Shift Measurements for Studying Ionospheric Fine Structure using Earth Satellites—M. C. Thompson, Jr, and D. M. Waters. (PROC. IRE, vol. 46, p. 1960; December, 1958.)

551.510.535 800

On the Electron Production Rate in the F_2 Region of the Ionosphere—S. Datta. (*Indian J. Phys.*, vol. 32, pp. 483–491; October, 1958.) The F_2 region, between "bottom" and height of maximum density, is divided into four equal columns, the mean production rate in each being calculated. The results show a diurnal variation of the rate with a single peak at about half an hour before noon.

551.510.535 801

Drift Observations Evaluated by the Method of 'Similar Fades'—E. Harnischmacher and K. Rawer. (*J. Atmos. Terr. Phys.*, vol. 13, pp. 1–16; December, 1958.) A regular interference model is considered as opposed to the usual purely random model. Some of the observed features are well explained by a model which lies between these two views, provided a finite lifetime is assumed for the irregularities. Large changes of drift velocity can be explained by assuming small vertical velocities.

551.510.535 802

Bifurcations in the F Region at Baguio, 1952–1957—V. Marasigan. (*J. Atmos. Terr. Phys.*, vol. 13, pp. 26–31; December, 1958.) A five-year statistical survey of bifurcations at Baguio, Phillipines is presented, and it is shown that the condition for bifurcation is mainly governed by the parameters h_m and y_m of the F_2 layer. h_m depends on latitude whilst y_m depends on solar activity.

551.510.535 803

The Diurnal Variation of f_0F_2 near the Auroral Zone during Magnetic Disturbances—B. Maehlum. (*J. Atmos. Terr. Phys.*, vol. 13, pp. 187–190; December, 1958.)

551.510.535 804

Horizontal Drifts and Temperature in the Lower Part of Region E—W. J. G. Beynon and G. L. Goodwin. (*J. Atmos. Terr. Phys.*, vol. 13, pp. 180–182; December, 1958.) Drift velocities deduced from the fading of CW signals at oblique incidence are related to results obtained using other techniques. A minimum velocity at 80–90 km is indicated and a possible connection with a temperature minimum at the same height is discussed.

551.510.535 805

Height Gradient of Electron Loss in the F Region—V. Marasigan. (*J. Atmos. Terr. Phys.*, vol. 13, pp. 107–112; December, 1958.) A theoretical expression is derived for an exponential height gradient of the electron loss coefficient in the F region on the assumption that this gradient completely accounts for F_1–F_2 bifurcation. Five models are investigated; they are second-power, linear, cosine, parabolic, and quasi-parabolic distributions of electron density with height.

551.510.535 806

The Electron Distribution in the Ionosphere over Slough: Part 2—Disturbed Days—J. O. Thomas and A. Robbins. (*J. Atmos. Terr. Phys.*, vol. 13, pp. 131–139; December, 1958.) The results are analysed of the distribution with height for three months in a year of low, and three months in a year of high sunspot number. It is shown that the variation in $h'F_2$ is not a quantitative guide to the changes in height of the F_2 layer during storms. The electron distributions for five individual storms are thoroughly investigated. Some important ionospheric changes have been noticed during a world-wide sudden impulse. Part 1: 1734 of 1958 (Thomas et al.).

551.510.535 807

On Instrument Effects in Ionosphere Data—H. J. Albrecht. (*J. Atmos. Terr. Phys.*, vol. 13, pp. 173–175; December, 1958.)

551.510.535:523.164.4 808

Abnormal Ionospheric Behaviour at 10 Metres Wavelength—M. Krishnamurthi, G. S. Sastry, and T. S. Rao. (*Curr. Sci.*, vol. 27, pp. 332–333; September, 1958.) During observations at Hyderabad, India, of cosmic radio noise total reflection of cosmic noise was observed on three occasions near sunrise. The effect is assumed to be caused by a locally high concentration of matter in the upper ionospheric layers.

551.510.535:621.396.11 809

On the Approximate Daytime Constancy of the Absorption of Radio Waves in the Lower Ionosphere—S. Chapman and K. Davies. (*J. Atmos. Terr. Phys.*, vol. 13, pp. 86–89; December, 1958.) The lower part of the D layer is due to photo-detachment of electrons from negative ions the concentration of which is large and nearly constant from day to night. The optical depth is small for the photo-detachment radiation but large for the radiation which can ionize neutral particles.

551.510.535:621.396.11 810

Ionospheric Absorption over Delhi—B. V. T. Rao and M. K. Rao. (*J. Instn Telecommun. Engrs, India*, vol. 4, pp. 205–208; September, 1958.) An analysis of further measurements made at 5 mc until October 1957. See also 3461 of 1958 (Mazumdar).

551.510.535:621.396.11:621.396.677.4 811

A New Antenna to Eliminate Ground-Wave Interference in Ionospheric Sounding Experiments—R. S. Macmillan, W. V. T. Rusch, and R. M. Golden. (*J. Atmos. Terr. Phys.*, vol. 13, pp. 183–186; December, 1958.)

551.594.5:621.396.11 812

Radio Reflections on Low Frequencies from 75–90 km Height during Intense Aurora Activity—W. Stoffregen. (*J. Atmos. Terr. Phys.*, vol. 13, pp. 167–169; December, 1958.)

551.594.6 813

Simultaneous Recording of Atmospherics on Four Different Frequency Bands in the Low-Frequency Region—M. W. Chiplonkar and V. N. Athavale. (*J. Atmos. Terr. Phys.*, vol. 13, pp. 32–37; December, 1958.) The number and intensity of atmospherics are recorded simultaneously on four narrow bands at 85, 125, 175, and 455 kc. Results show atmospherics with large field strengths to be less frequent than those with small field strengths on all bands, and the field strength of atmospherics to be a maximum in the 125 kc band.

551.594.6:551.594.221 814

The Relationship between Atmospheric Radio Noise and Lightning—F. Horner. (*J. Atmos. Terr. Phys.*, vol. 13, pp. 140–154; December, 1958.) In Europe atmospherics in a bandwidth of 300 c at 10 kc have median amplitude, amplitude range, and frequency occurrence in accordance with that expected from lightning discharges to the ground. In Australia atmospherics from tropical storms were found to consist of numerous pulses the largest of which probably originated in ground strokes. The origin of the smaller pulses is not clear. Some observations on atmospherics in the HF band are also described.

551.594.6:621.3.087.4/.5 815
Waveforms of Atmospherics with Superimposed Pulses Recorded with an Automatic Atmospherics Recorder—B. A. P. Tantry and R. S. Srivastava. (*J. Atmos. Terr. Phys.*, vol. 13, pp. 38–42; December, 1958.) The recorder design is briefly outlined and observations are discussed in which "stepped" pulses from one lightning discharge are superimposed on the waveform from a second discharge. See also 3824 of 1958 (Tantry).

LOCATION AND AIDS TO NAVIGATION

621.396.932.1 816
A New Method of Component Determination in Radio Direction-Finding for Coherent Waves—H. Gabler and M. Wächtler. (*Elektrotech. Z.*, vol. 79, pp. 383–388; June 1, 1958.) The three ellipses produced by three suitably spaced crossed-loop antennas with a two-channel cathode-ray DF equipment are superimposed on the screen. Bearings free from night effect can be obtained even under unfavourable site conditions.

621.396.933.1 817
New V.H.F. Direction-Finding Equipment—(*Brit. Commun. Electronics*, vol. 5, p. 681; September, 1958.) A brief description is given of automatic equipment using a rotating Adcock antenna. The operational range is about 100 miles for aircraft flying at 10,000 ft and radiating 5W.

621.396.96 818
Radar Systems with Electronic Sector Scanning.—D. E. N. Davies—(*J. Brit. IRE*, vol. 18, pp. 709–713; December, 1958.) The application to radar of a system previously described in relation to underwater acoustic echo-ranging [3676 of 1958 (Tucker et al.)] is discussed. Information rate is much higher than that of a conventional radar system.

621.396.96:551.51 819
Radar Echoes from Atmospheric Inhomogeneities—R. F. Jones. (*Quart. J. R. Met. Soc.*, vol. 84, pp. 437–442; October, 1958.) A quantitative examination of the possibility of radar echoes being caused by scattering or reflection from atmospheric inhomogeneities shows this to be theoretically possible but necessitating large changes in refractive index.

621.396.962.33 820
Decca Doppler and Airborne Navigation—T. Gray and M. J. Moran. (*Brit. Commun. Electronics*, vol. 5, pp. 764–771; October, 1958.) The techniques of transmission and reception used in Doppler systems are examined. The Decca Doppler sensor uses a self-coherent pulsed "Janus" system with a symmetrical four-beam aerial configuration. A qualitative description of the Decca Integrated Airborne Navigation system (DIAN) is given.

621.396.963 821
Accurate Method for Correction of Slant-Range Distortion in High-Altitude Radars and a Contribution to the Optics of Reflecting Conical Surfaces—L. Levi. (*J. Opt. Soc. Amer.*, vol. 48, pp. 680–686; October, 1958.)

621.396.967:621.396.65 822
Microwave Links for Radar Networks—Sutherland. (See 976.)

621.396.969.33:621.396.677.7 823
Slotted-Waveguide Array for Marine Radar—Byers and Katchky. (See 711.)

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215 824
Internal Photoeffect and Exciton Diffusion in Cadmium and Zinc Sulphides—M. Balkanski and R. D. Waldron. (*Phys. Rev.*, vol. 112, pp. 123–135; October 1, 1958.) The absorption spectra of CdS and ZnS and the photoconductivity produced by illumination at a distance from the electrodes were studied to determine the mechanisms of photo-electron interaction and energy transport in these materials.

535.215 825
Photoemissive, Photoconductive, and Optical Absorption Studies of Alkali-Antimony Compounds—W. E. Spicer. (*Phys. Rev.*, vol. 112, pp. 114–122; October 1, 1958.) Experimental methods and results are given. Values for the band gaps and electron affinities are tabulated, and the conductivity types as indicated by photo-emission data are listed.

535.215:546.47:31 826
Photo-properties of Zinc Oxide with Ohmic and Blocking Contacts—H. J. Gerritsen, W. Ruppel, and A. Rose. (*Helv. Phys. Acta*, vol. 30, pp. 504–512; November 20, 1957.) Report of measurements of primary and secondary photocurrents in fine-grain ZnO layers.

535.215+535.37]:546.482.21 827
The Mechanism of Energy Transfer in Cadmium Sulphide Crystals—I. Broser and R. Broser-Warminsky. (*J. Phys. Chem. Solids*, vol. 6, pp. 386–400; In German.) The transfer of the energy of excitation from the point of absorption to non-irradiated regions over relatively large distances may be explained by the scattering and reabsorption of the incident light or the luminescent radiation generated in the crystal. The hypothesis of energy conduction by excitons is not confirmed.

535.215:546.482.21 828
Photoconductivity and Crystal Size in Evaporated Layers of Cadmium Sulphide—J. M. Gilles and J. Van Cakenberghe. (*Nature, London*, vol. 182, pp. 862–863; September 27, 1958.) The effect of heating a layer of CdS crystals in contact with a film of evaporated Ag to 500°–600°C in an inert atmosphere is discussed.

535.215:546.482.21 829
Progress in Cadmium Sulphide—L. L. Antes. (IRE TRANS. ON COMPONENT PARTS, vol. CP-4, pp. 129–132; December, 1957. Abstract, PROC. IRE, vol. 46, p. 801; April, 1958.)

535.215:546.817.231:539.23 830
Photoconductivity in Chemically Deposited Films of Lead Selenide—D. H. Roberts and J. E. Baines. (*J. Phys. Chem. Solids*, vol. 6, pp. 184–189; August, 1958.)

535.37 831
Electrophotoluminescent Amplification—R. E. Halsted. (*J. Appl. Phys.*, vol. 29, pp. 1706–1708; December, 1958.) An alternating electric field applied parallel to an excitation gradient in some sulphide phosphors can produce a pronounced characteristic modulation of the photoluminescent emission. Experimental results on this effect are discussed.

535.376:546.472.21 832
Electroluminescence of Zinc Sulphide Phosphors as an Equilibrium Process—W. Lehmann. (*J. Opt. Soc. Amer.*, vol. 48, pp. 647–653; September, 1958.) Predictions of phosphor characteristics based on the equilibrium condition relating monomolecular collision excitation and bimolecular recombination processes are compared with experimental results.

535.376:546.472.21 833
The Significance of Boundary Layers and Polarization Fields for Electroluminescence—D. Hahn and F. W. Seemann. (*Z. Phys.*, vol. 149, pp. 486–503; October 31, 1957.) Luminescence of ZnS-based phosphors was investigated using excitation by square pulses of alternating or single polarity. See also 2156 of 1957.

535.376:546.472.21 834
Cathodo-electroluminescence Phenomena in ZnS Phosphors—H. Gobrecht, H. E. Gumlich. H. Nelkowski, and D. Langer. (*Z. Phys.*, vol. 149, pp. 504–510; October 31, 1957.) Two powder phosphors, one electroluminescent and the other nonelectro luminescent but with pronounced enhancement effect, are investigated.

535.376:546.472.21 835
Electroluminescence of ZnS Single Crystals with Cathode Barriers—D. R. Frankl. (*Phys. Rev.*, vol. 111, pp. 1540–1549; September 15, 1958.) Certain ZnS crystals show electroluminescence predominantly near the cathode. The major emission from Cu-activated crystals occurs as a burst of light when the exciting voltage is suddenly removed, the burst being quenched by a voltage in the initial direction or enhanced by a voltage in the opposite direction. These results, as well as the emission peaks obtained under sinusoidal voltage excitation, are explained in terms of ionization of luminescent centers in a barrier region.

537.226/.227 836
Dielectric and Thermal Study of $(NH_4)_2SO_4$ and $(NH_4)_2BeF_4$ Transitions—S. Hoshino· K. Vedam, Y. Okaya, and R. Pepinsky. (*Phys. Rev.*, vol. 112, pp. 405–412; October 15, 1958.)

537.226 837
Electrical Dispersion Phenomena in Inhomogeneous Dielectrics—R. Parker and M. S. Smith. (*J. Electronics Control*, vol. 5, pp. 354–361; October, 1958.) The theory of Koops (162 of 1952) is generalized to include an arbitrary distribution of layer thicknesses and orientations. Observed deviations from the dispersion equations are ascribed to inhomogeneities in the current density vector. The evaluation of surface-layer parameters is discussed.

537.226 838
The Ternary Systems $BaO-TiO_2-SnO_2$ and $BaO-TiO_2-ZrO_2$—G. H. Jonker and W. Kwestroo. (*J. Amer. Ceram. Soc.*, vol. 41, pp. 390–394; October 1, 1958.)

537.226:621.315.612 839
Lightweight Ceramic Materials as High-Frequency Dielectrics—J. L. Pentecost and P. E. Ritt. (IRE TRANS. ON COMPONENT PARTS vol. CP-4, pp. 133–135; December, 1957. Abstract, PROC. IRE, vol. 46, p. 801; April, 1958.)

537.227 840
Room-Temperature Ferroelectricity in Lithium Hydrazinium Sulphate, $Li(N_2H_5)SO_4$—R. Pepinsky, K. Vedam, Y. Okaya, and S. Hoshino. (*Phys. Rev.*, vol. 111, pp. 1467–1468; September 15, 1958.) A method of producing (and protecting) large crystals is described together with details of an investigation of the ferroelectric behaviour and crystallographic structure.

537.227 841
Ammonium Hydrogen Sulphate: A New Ferroelectric with Low Coercive Field—R. Pepinsky, K. Vedam, S. Hoshino, and Y. Okaya. (*Phys. Rev.*, vol. 111, pp. 1508–1510; September 15, 1958.) Details are given of the ferroelectric behaviour of $(NH_4)HSO_4$ in the range −3°C to −119°C, together with details of the crystallographic structure. A particular feature is the low coercive field which is about 150 v/cm at −13°C.

537.227:539.2 842
Polarization Fluctuations in a Ferroelectric Crystal—R. E. Burgess. (*Can. J. Phys.*, vol. 36,

pp. 1569–1581; November, 1958.) The theory of thermal fluctuations of electrical polarization in a ferroelectric crystal is considered, with special attention to temperatures in the neighbourhood of the Curie point. By thermodynamical analysis of a dipole model, the fluctuations and their spectral density are related respectively to the real and imaginary parts of the susceptibility of the crystal.

537.227:546.431.824–31 843
Transition to the Ferroelectric State in Barium Titanate—D. Meyerhofer. (*Phys. Rev.*, vol. 112, pp. 413–423; October 15, 1958.) Besides raising the cubic-tetragonal transition about 15°C with an electric field along the cube-edge direction, an orthorhombic phase was induced above the Curie point by a field along a face-diagonal direction. Birefringence, polarization, and dielectric constant were measured above and below the Curie point as functions of field strength and field direction.

537.227:546.431.824–31 844
Polarization Changes during the Process of Ageing in Ferroelectrics of the $BaTiO_3$-Type—Z. Pajak and J. Stankowski. (*Proc. Phys. Soc.*, vol. 72, pp. 1144–1146; December 1, 1958.) The dielectric-hysteresis loops of barium metatitanate ceramics become constricted with age and eventually form a narrow double loop. The ageing process may be reversed by heating above the Curie point.

537.228.1 845
Some Piezoelectric Properties of Polycrystalline Solid Solutions (Ba, Sr)TiO_3, Ba(Ti, Sn)O_3 and Ba(Ti, Zr)O_3—V. A. Bokov. (*Akust. Z.*, vol. 3, pp. 104–108; April/June, 1957.) In certain solid solutions, the electromechanical response and effective piezoelectric modulus depend on the voltage of the polarizing field and temperature. This response and coefficient pass through a maximum, due to domain orientation, particularly in solid solutions of the type Ba(Ti, Sn)O_3 and Ba(Ti, Zr)O_3.

537.228.1:546.472.21 846
Theory of the Piezoelectric Effect in the Zinoblende Structure—J. L. Birman. (*Phys. Rev.*, vol. 111, pp. 1510–1514; September 15, 1958.) A theory is developed which leads to an equation relating the macroscopic piezoelectric constant to the static and dynamic effective charges, and a lattice parameter which relates the internal strain to the external strain. An order-of-magnitude check of the theory is made on ZnS.

537.311.33 847
The Prediction of Semiconducting Properties in Inorganic Compounds—C. H. L. Goodman. (*J. Phys. Chem. Solids*, vol. 6, pp. 305–314; September, 1958.) Various criteria are presented for predicting possible semiconductor behavior in inorganic compounds, based on the requirements of saturated ionic-covalent bonding. It is shown that new series of semiconducting compounds can be derived from known ones by replacing one element by pairs from other groups of the periodic table, while keeping the valence-electron:atom ratio constant. These concepts are illustrated with particular reference to diamond-type lattices.

537.311.33 848
Calculations on the Shape and Extent of Space-Charge Regions in Semiconductor Surfaces—G. C. Dousmanis and R. C. Duncan, Jr. (*J. Appl. Phys.*, vol. 29, pp. 1627–1629; December, 1958.) Curves showing the potential as a function of distance inside a semiconductor are obtained by numerical integration of the Poisson equation. The results apply to all semiconductors.

537.311.33 849
The Dependence of Minority-Carrier Lifetime on Majority-Carrier Density—D. M. Evans. (PROC. IRE, vol. 46, pp. 1962–1963; December, 1958.) The minority-carrier lifetime in Ge and Si is inversely proportional to the majority-carrier density only if the recombination levels lie near the appropriate band edge and/or the semiconductor is only weakly extrinsic.

537.311.33 850
Influence of Crystal Lattice Vibrations on the Production of Electron-Hole Pairs in a Strong Electric Field—L. V. Keldysh. (*Zh. Eksp. Teor. Fiz.*, vol. 34, pp. 962–968; April, 1958.) The probability for the production of an electron-hole pair in a semiconductor is calculated considering electron-phonon interaction. Other processes which may influence the diffusion of a valence electron into the conduction band are considered, in particular, the absorption of several phonons which may be decisive in relatively weak fields.

537.311.33 851
Semiconductor Surface Potential and Surface States from Field-Induced Changes in Surface Recombination—G. C. Dousmanis. (*Phys. Rev.*, vol. 112, pp. 369–380; October 15, 1958.) The variations in surface recombination are detected by changes in the reverse current of large-area "back-surface" diodes. Observation of a maximum of surface recombination in terms of applied field provides a reference point from which the zero-field value of the surface potential can be evaluated, and the dependence of the surface recombination velocity on the surface potential can be established.

537.311.33 852
Junction Capacitance and Related Characteristics using Graded Impurity Semiconductors—L. J. Giacoletto. (IRE TRANS. ON ELECTRON DEVICES, vol. ED-4, pp. 207–215; July, 1957. Abstract, PROC. IRE, vol. 45, p. 1760; December, 1957.)

537.311.33 853
The *K*-Edge Structures of the Elements of Various $A_{III}B_V$ Compounds—W. Eberbeck. (*Z. Phys.*, vol. 149, pp. 412–424; October 31, 1957. The fine structure of the K X-ray absorption spectrum was investigated for Ga in GaP, GaAs, and GaSb, for As in GaAs and InAs, for Zn in ZnS, and for pure Ge.

537.311.33:537.32 854
The Effect of Strain on the Thermoelectric Properties of a Many-Valley Semiconductor—J. R. Drabble. (*J. Electronics Control.* vol. 5, pp. 362–372; October, 1958.) A general treatment is given for non degenerate materials. It is shown that the absence of effects other than those due to carrier transfer between valleys can be checked by the condition that the trace of the conductivity tensor is unaltered by the strain. Under these conditions, the conductivity and thermoelectric power tensors can be combined to give an expression involving only the relative shifts in the minimum energies of the valleys.

537.311.33:538.632 855
Hall Effect in Semiconductor Compounds—M. J. O. Strutt. (*Electronic Radio Engr.*, vol. 36, pp. 2–10; January, 1959.) The effect in InAs and InSb, and the influence of probe shape and position are discussed. Applications to several types of wattmeter, to oscillators, to a flux-density meter, and to a receiver mixer stage are described.

537.311.33:538.632 856
Hall Mobility of Carriers in Impure Nondegenerate Semiconductors—M. S. Sodha and P. C. Eastman. (*Phys. Rev.*, vol. 112, p. 44; October 1, 1958.) The case of low temperatures and high impurity concentrations is considered.

537.311.33:[546.28+546.289 857
Lattice Vibrational Spectra of Si and Ge—H. Cole and E. Kineke. (*Phys. Rev. Lett.*, vol. 1, pp. 360–361; November 15, 1958.)

537.311.33:[546.28+546.289 858
Theoretical Surface Conductivity Changes and Space Charge in Germanium and Silicon—V. O. Mowery. (*J. Appl. Phys.*, vol. 29, pp. 1753–1757; December, 1958.) Graphs are given showing the surface conductivity and space charge as functions of resistivity for various values of surface potential.

537.311.33:546.28 859
The Measurement of Surface Recombination Velocity on Silicon—A. H. Benny and F. D. Morten. (*Proc. Phys. Soc.*, vol. 72, pp. 1007–1012; December 1, 1958.) The surface recombination velocity on a Si filament can be determined by the measurement of the spectral distribution of the photoconductivity. On n-type Si the surface recombination velocity can be varied from less than 75 to greater than 7500 cm. sec^{-1} by changing the ambient gas.

537.311.33:546.28 860
Microwave Spin Echoes from Donor Electrons in Silicon—J. P. Gordon and K. D. Bowers. (*Phys. Rev. Lett.*, vol. 1, pp. 368–370; November 15, 1958.) The relaxation time in Si containing Li and P has been measured at a frequency of 23 kmc using a heterodyne paramagnetic spectrometer. Results indicate that a spin-echo device could provide a storage system in which each element could store more than 10^4 bits of information.

537.311.33:546.28 861
Neutron-Bombardment Damage in Silicon—G. K. Wertheim. (*Phys. Rev.*, vol. 111, pp. 1500–1505; September 15, 1958.) Neutron-bombardment damage in Si is compared to electron-bombardment effects which have been previously analysed (3515 of 1958). A discrete energy level 0.27 ev above the valence band is produced by both neutrons and electrons. A spectrum of energy levels running from 0.16 ev below the conduction band toward the middle of the gap is ascribed to a defect pair with variable spacing.

537.311.33:546.28 862
Recombination Properties of Gold in Silicon—G. Bemski. (*Phys. Rev.*, vol. 111, pp. 1515–1518; September 15, 1958.) The capture of electrons in p-type Si occurs through the Au donor le^el with a capture cross-section of $3.5 \times 10^{-3}cm^{-15}$ at 300°K; this capture cross-section varies as $T^{-2.5}$ between 200° and 500°K. In n-type Si the electron capture cross-section is $5\times 10^{-16}cm^2$ and is of independent temperature; the hole capture cross-section is 1×10^{-15} cm^2 at 300°K and varies as T^{-4}. The capture in this case occurs through the Au acceptor level.

537.311.33:546.28:535.215 863
Quantum Yield of Photoionization in Silicon—V. S. Vavilov and K. I. Britsyn. (*Zh. Eksp. Teor. Fiz.*, vol. 34, pp. 1354–1355; May, 1958.) Results of an investigation of the photo-effect in p-n junctions obtained by thermal diffusion of phosphorus indicate an increase in quantum yield and the presence of impact ionization by liberated carriers.

537.311.33:[546.28+546.289]:535.39–15 864
Antireflection Coatings for Germanium and Silicon in the Infrared—J. T. Cox and G. Hass. (*J. Opt. Soc. Amer.*, vol. 48, pp. 677–680; October, 1958.)

537.311.33:546.289 865
Production of Dislocations in Germanium by Thermal Shock—R. S. Wagner. (*J. Appl. Phys.*, vol. 29, pp. 1679–1682; December, 1958.) The results indicate that the imperfections are not generated during growth, but occur afterwards, as a result of thermal shock when the growth is terminated.

537.311.33:546.289 866
Effect of Monoenergetic Fast Neutrons on n-Type Germanium—S. L. Ruby, F. D. Schupp, and E. D. Wolley. (*Phys. Rev.*, vol. 111, pp. 1493–1496; September 15, 1958.) The density of vacancy-interstitial pairs produced unit of fast-neutron flux on n-type Ge has been measured experimentally using monoenergetic neutrons. The observed changes in resistivity and Hall coefficients indicate charge carrier removal rates per unit of neutron flux much lower than that calculated from current models.

537.311.33:546.289 867
Photoconductive Response of Single-Crystal Germanium Layers prepared by the Pyrolytic Decomposition of GeI_2—D. C. Cronemeyer. (*J. Appl. Phys.*, vol. 29, pp. 1730–1735; December, 1958.) The photo-response measured at 77°K was found to extend to about 6 μ, in agreement with the thermal activation energy. Doping of the Ge with Au and Ag was attempted, but no positive identification of the doping agents was obtained from the photoconductive response.

537.311.33:546.289 868
Radiation-Induced Recombination Centers in Germanium—O. L. Curtis, Jr, J. W. Cleland, and J. H. Crawford, Jr. (*J. Appl. Phys.*, vol. 29, pp. 1722–1729; December, 1958.) The effect of energetic particle bombardment on the minority carrier lifetime has been measured. It is concluded that a defect state located 0.20 ev below the conduction band dominates the recombination process.

537.311.33:546.289 869
Radiative Recombination in Germanium—P. H. Brill and R. F. Schwarz. (*Phys. Rev.*, vol. 112, pp. 330–333; October 15, 1958.) Dependence on excess-carrier density and on equilibrium-carrier density was studied by simultaneous measurements of output radiation and of photoconductivity as functions of incident light intensity. Results confirm the theory.

537.311.33:546.289 870
Magnetoresistance in n-Type Germanium at Low Temperatures—R. A. Laff and H. Y. Fan. (*Phys. Rev.*, vol. 112, pp. 317–321; October 15, 1958.) The effective anisotropy parameter K decreased from $\sim$20 at 300°K to $\sim$6 at 20°K, but values near 20 were again obtained at 7°K and 4.2°K. Introduction of compensating acceptors showed that the decrease of K was due to anisotropic scattering by ionized impurities. Below 7°K, scattering was controlled by neutral impurities with essentially isotropic relaxation time.

537.311.33:546.289 871
Infrared Absorption by Conduction Electrons in Germanium—H. J. G. Meyer. (*Phys. Rev.*, vol. 112, pp. 298–308; October 15, 1958.) A theory is developed taking into account the multivalley structure of the condution band and present knowledge of the scattering mechanism. Explicit calculations are possible at all relevant wavelengths and temperatures.

537.311.33:546.289 872
Oscillation of the Electrical Resistance of n-Type Germanium in Strong Pulsed Magnetic Fields—I. G. Fakidov and E. A. Zavadskiĭ. (*Zh. Eksp. Teor. Fiz.*, vol. 34, pp. 1036–1037; April, 1958.) Note on measurements of the resistance variation in single-crystal Ge in a transverse pulse magnetic field up to 120 kg at temperatures 300, 77 and 20°K.

537.311.33:546.289 873
Surface Effects in Electron-Irradiated Ge at 80°K—W. E. Spear. (*Phys. Rev.*, vol. 112, pp. 362–369; October 15, 1958.) Results are given of photoconductivity, surface-conductance, Hall-, and field-effect measurements made before and after electron irradiation. Use of both low- and high-energy irradiation distinguishes between volume and surface effects. Surface rather than volume effects are responsible for photoconductivity changes beyond the fundamental absorption edge.

537.311.33:546.289:535.215 874
Delayed Electron Emission and External Photoeffect of Germanium after Electron Bombardment—K. Seeger. (*Z. Phys.*, vol. 149, pp. 453–470; October 31, 1957.) Emission was investigated in the temperature range 103°–670°K. An interpretation of the effect as being due to slow surface states is discussed.

537.311.33:546.289:538.569 875
Experimental Evidence for Carriers with Negative Mass—G. C. Dousmanis, R. C. Duncan, Jr, J. J. Thomas, and R. C. Williams. (*Phys. Rev. Lett.*, vol. 1, pp. 404–407; December 1, 1958.) Cyclotron resonance experiments have been carried out on Ge crystals at 4°K. A spectrum of heavy hole has been revealed which may be assigned to carriers of negative mass.

537.311.33:546.289.231 876
The Crystal Structure of Germanium Selenide GeSe—A. Okazaki. (*J. Phys. Soc. Japan*, vol. 13, pp. 1151–1155; October, 1958.)

537.311.33:546.47–31 877
An Approach to Intrinsic Zinc Oxide—W. Ruppel, H. J. Gerritsen, and A. Rose. (*Helv. Phys. Acta.* vol. 30, pp. 495–503; November 20, 1957.) Measurements have been made on layers of finely divided ZnO powder the resistivity of which ($10^{17}\,\Omega\cdot$ cm) is closer to that of intrinsic ZnO than that of single crystals. The formation of depletion layers and the rectifying action observed with ohmic and blocking contacts are discussed.

537.311.33:546.47–31 878
The Preparation of Zinc Oxide Single Crystals with Definied Impurity Content—G. Bogner and E. Mollwo. (*J. Phys. Chem. Solids*, vol. 6, pp. 136–143; August, 1958. In German.) A method is described by which ZnO crystals of a definite composition can be deposited from the gas phase. Measurements of the composition, conductivity, and infrared absorption of crystals doped with Cu or In are reported.

537.311.33:546.47–31 879
The Concentration and Mobility of Electrons in Zinc Oxide Single Crystals with Defined Impurity Content—H. Rupprecht. (*J. Phys. Chem. Solids*, vol. 6, pp. 144–154; August, 1958. In German.) The conductivity and Hall effect in synthetic crystals of doped ZnO have been measured in tce temperature range 65–700°K. Results are similar to those for other semiconductors such as Ge and Si.

537.311.33:546.47–31 880
Field Effect and Photoconductivity in ZnO Single Crystals—G. Heiland. (*J. Phys. Chem. Solids*, vol. 6, pp. 155–168; August, 1958.) Report and discussion of the results of measurements made on ZnO crystals with surface conductivity varying over a wide range.

537.311.33:546.681.19 881
The Preparation and Properties of Gallium Arsenide Single Crystals—J. M. Whelan and G. H. Wheatley. (*J. Phys. Chem. Solids*, vol. 6, pp. 169–172; August, 1958.) Apparatus is described for the preparation and purification of GaAs by horizontal zone refining, and for the production of single crystals by the floating-zone method. Resistivities up to $10^6\,\Omega\cdot$cm have been obtained. Hall coefficient and conductivity data for n-type samples are given. Electron mobility, the energy gap, and the position of an impurity level are deduced.

537.311.33:546.681.19 882
Diffusion, Solubility, and Electrical Behaviour of Copper in Gallium Arsenide—C. S. Fuller and J. M. Whelan. (*J. Phys. Chem. Solids*, vol. 6, pp. 173–177; August, 1958.)

537.311.33:546.682.18 883
Optical Properties of n-Type InP—R. Newman. (*Phys. Rev.*, vol. 111, pp. 1518–1521; September 15, 1958.)

537.311.33:546.682.19:537.312.9 884
Piezoresistance Constants of n-Type InAs—A. J. Tuzzolino. (*Phys. Rev.*, vol. 112, p. 30; October 1, 1958.) Results of measurements from 77°K to 300°K are consistent with a spherical conduction-band model.

537.311.33:546.72.23 885
$FeSe_2$, a Semiconductor containing Iron—G. Fischer. (*Can. J. Phys.*, vol. 36, pp. 1435–1438; October, 1958.) The preparation of this compound is described; the results of electrical-resistivity and Hall-coefficient measurements and X-ray examination are summarized.

537.311.33:546.812.221 886
The Semiconductor Properties of Synthetic Herzenbergite (SnS) Crystals—H. Gobrecht and A. Bartschat. (*Z. Phys.*, vol. 149, pp. 511–522; October 31, 1957.)

537.311.33:546.824–31:537.312.9 887
Piezoresistivity in Reduced Single-Crystal Rutile (TiO_2)—L. E. Hollander, Jr. (*Phys. Rev., Lett.*, vol. 1, pp. 370–371; November 15, 1958.) The effect of stress on resistivity has been measured at different temperatures.

537.311.33:621.314.63 888
Alloy Junctions in Semiconducting Devices—D. F. Taylor. (*Research. London*, vol. 11, pp. 335–338; September, 1958.) Outline of theory on which methods of preparing alloy junctions are based.

537.311.33:621.318.57 889
Current Build-Up in Semiconductor Devices—W. Shockley and J. Gibbons. (PROC. IRE, vol. 46, pp. 1947–1949; December, 1958.) Assuming that the minority-carrier densities and their associated currents have an exponential rise during the switching action, a solution to the partial differential equations can be found from which the efficiencies of various designs of switching devices may be derived.

538.22 890
An Interpretation of the Magnetic Properties of the Perovskite-Type Mixed Crystals $La_{1-x}Sr_xCoO_{3-\lambda}$—J. B. Goodenough. (*J. Phys. Chem. Solids*, vol. 6, pp. 287–297; August, 1958.)

538.221 891
Interatomic Distances in Ferromagnetics—F. M. Gal'perin. (*Zh. Eksp. Teor. Fiz.*, vol. 34, pp. 1000–1003; April, 1958.)

538.221 892
The Change of Spontaneous Magnetization with Hydrostatic Pressure—D. Gugan. (*Proc. Phys. Soc.*, vol. 72, pp. 1013–1026; December 1, 1958.)

538.221 893
Dipolar Energy. Application to the Magnetostatic Energy of α-Fe_2O_3—F. Bertaut. (*Compt. Rend. Acad. Sci., Paris*, vol. 246, pp. 3335–3337; June 16, 1958.) The magnetic anisotropy and weak ferromagnetism of α-Fe_2O_3 cannot be explained in terms of dipolar energy.

538.221 894
Sendust Flake—a New Magnetic Material for Low-Frequency Application—W. M. Hubbard, E. Adams, and J. F. Haben. (IRE TRANS. ON COMPONENT PARTS, vol. CP-4, pp. 2–6; March, 1957. Abstract, PROC. IRE, vol. 45, p. 1035; July, 1957.)

538.221:539.231 895
Magnetic, Electrical and Electron-Optical Investigations of the Thermal Transformation of fathode-Sputtered Nickel films—L. Reimer. (*Z. Phys.*, vol. 149, pp. 425–431; October 31, 1957.)

538.221:621.3.042.2 896
Cube-Oriented Magnetic Sheet—(*J. Metals, N.Y.*, vol. 10, pp. 507–511; August, 1958.)
A Major Advance in Magnetic Materials—G. W. Wiener and K. Detert (pp. 507–508).
Magnetic Properties of Cube-Textured Transformer Sheet—J. L. Walter, W. R. Hibbard, Jr, H. C. Fiedler, H. E. Grenoble, R. H. Pry and P. G. Frischmann (pp. 509–511).

538.221:[621.318.124+621.318.134 897
Data on Ferrite Core Materials—A. C. Hudson and E. J. Stevens. (*Electronic Eng.*, vol. 30, pp. 718–719; December, 1958.) Charts are presented which permit comparison of the permeability and loss at low levels for various ferrites over the frequency range 100 kc–100 mc.

538.221:[621.318.124+621.318.134 898
Effect of Neutron Irradiation on the Curie Temperature of a Variety of Ferrites—E. I. Salkovitz, G. C. Bailey, and A. I. Schindler. (*J. Appl. Phys.*, vol. 29, pp. 1747–1748; December, 1958.) No significant change in Curie temperature after irradiation was found for either magnetically soft or hard ferrites.

538.221:621.318.124 899
Magnetization Studies and Possible Magnetic Structure of Barium Ferrate III—W. E. Henry. (*Phys. Rev.*, vol. 112, pp. 326–327; October 15, 1958.)

538.221:621.318.134 900
Modified Rotational Model of Flux Reversal—E. M. Gyorgy. (*J. Appl. Phys.*, vol. 29, pp. 1709–1712; December, 1958.) The mechanism of flux reversal in square-loop ferrites is analyzed showing that the flux in a toroid can be reversed by rotation without prohibitively large demagnetizing fields. This agrees with observations. See also 534 of 1958.

38.221:621.318.134 901
Magnetic Anisotropy Constant of Yttrium-Iron Garnet at 0°K—B. R. Cooper. (*Phys. Rev.*, vol. 112, pp. 395–396; October 15, 1958.)

538.221:621.318.134 902
Low-Temperature Transition of Magnetic Anisotropy in Nickel-Iron Ferrite—N. Menyuk and K. Dwight. (*Phys. Rev.*, vol. 112, pp. 397–405; October 15, 1958.) A study of the magnetic properties of single-crystal samples of Ni-Fe ferrite has revealed an abrupt transition in the magnetic anisotropy characteristics at 10°K. Predictions based on a proposed model are in accord with experimental findings.

538.221:621.318.134 903
An Investigation into the Magnesium and Magnesium-Manganese Ferrite System—L. C. F. Blackman. (*J. Electronics Control*, vol. 5, pp. 373–384; October, 1958.) "The influence of both manganese and firing conditions on certain physical and chemical properties of iron-deficient magnesium ferrite has been investigated. The results suggest that Fe^{2+} and Mn^{3+} can coexist, at least to a certain extent, in the solid state. It is also shown that the excess of magnesia, which is added to improve the microwave performance of the materials, plays an important role in the chemistry of the ferrite phase."

538.221:621.318.134 904
Ferrimagnetism in the System Na_2O-ZnO-Fe_2O_3—A. H. Mones and E. Banks. (*J. Phys. Chem. Solids*, vol. 6, pp. 267–270; August, 1958.)

538.221:621.318.134 905
Origin of Weak Ferromagnetism in Rare-Earth Orthoferrites—R. M. Bozorth. (*Phys. Rev. Lett.*, vol. 1, pp. 362–363; November 15, 1958.)

538.221:621.318.134:538.569.4 906
Ferromagnetic Resonance in Polycrystalline Ferrites with Large Anisotropy: Part 1—General Theory and Application to Cubic Materials with a Negative Anisotropy Constant—E. Schlömann. (*J. Phys. Chem. Solids*, vol. 6, pp. 257–266; August, 1958.)

538.221:621.318.134:538.569.4 907
Ferrimagnetic Resonance in Yttrium-Iron Garnet at Liquid Helium Temperatures—J. F. Dillon, Jr. (*Physl Rev.*, vol. 111, pp. 1476–1478; September 15, 1958.) Experiments performed on single crystal spheres at 24 kmc are described. The paper gives in some detail the low-temperature results on line width and the variation of the resonance field with crystal direction and temperature.

538.221:621.318.134:538.569.4 908
Magnetostatic Modes in Ferrimagnetic Spheres—J. F. Dillon, Jr. (*Phys. Rev.*, vol. 112, pp. 59–63; October 1, 1958.) Experiments on Y-Fe garnet show that the positions of the modes are not functions of crystal direction as previously reported. Specimens produced by tumbling procedures may deviate slightly from sphericity and in these experiments truly spherical samples were used. Evidence has been given of the effect of dielectric inhomogeneities in the neighborhood of the sample in a ferrimagnetic-resonance experiment.

538.222:538.569.4 909
Paramagnetic-Resonance Spectrum of Gadolinium in Single Crystals of Thorium Oxide—W. Low and D. Shaltiel. (*J. Phys. Chem. Solids*, vol. 6, pp. 315–323; September, 1958.) Measurements at 290° and 90°K at 3 cm λ give line positions and intensities which can be explained by a crystalline field of cubic symmetry. It is suggested that oxygen vacancies are randomly distributed throughout the crystal.

539.2:537.311.31 910
Electronic Band Structures of the Alkali Metals and of the Noble Metals and their α-Phase Alloys—M. H. Cohen and V. Heine. (*Advances Phys.*, vol. 7, pp. 395–434; October, 1958.)

621.315.616.96 911
Electrical Properties of Epoxy Resins—C. F. Pitt, B. P. Barth, and B. E. Godard.—(IRE TRANS. ON COMPONENT PARTS, vol. CP-4, pp. 110–113; December, 1957. Abstract, PROC. IRE, vol. 46, p. 801; April, 1958.)

MATHEMATICS

517.942:538.566 912
Method of Liouville Applied to Weber's Equation—R. Meynieux. (*Compt. Rend. Acad. Sci., Paris*, vol. 246, pp. 3208–3210; June 9, 1958.) Applications include the propagation of a plane em wave across an ionospheric layer.

517.942:538.66 913
Gamma-Function Approximations Applied to Solutions of Weber's Equation—R. Meynieux. (*Compt. Rend. Acad. Sci., Paris*, vol. 246, pp. 3312–3314; June 16, 1958.)

MEASUREMENTS AND TEST GEAR

621.3.018.41(083.74):529.786:525.35 914
Comparison of Astronomical Time Measurements with Atomic Frequency Standards—J. P. Blaser and J. De Prins. (*Nature, London*, vol. 182, pp. 859–860; September 27, 1958.) Comparisons have been made in Neuchatel between "atomic" time determined by integration of the frequency of a quartz clock calibrated against an NH_3 maser, and astronomical time determined by means of photographic zenith tube and astrolabe. Plotting UT1 against atomic time shows the variation in the rate of rotation of the earth, which is in agreement with the measurements of Essen *et al.* (3195 of 1958).

621.3.018.41(083.74):621.396.11 915
Comparison of an Ammonia Maser with a Caesium Atomic Frequency Standard—J. P. Blaser and J. Bonanomi. (*Nature, London*, vol. 182, p. 859; September 27, 1958.) Note of a comparison made in Neuchâtel by means of MSF standard-frequency transmissions during 1957 and 1958 of the frequencies of an NH_3 maser and the Cs resonator of the National Physical Laboratory. Two different methods were used: *a*) differentiation of the phase of MSF time signals at 5 and 10 mc; *b*) direct frequency comparison using the 60-kc transmission. See also 1208 of 1958 (Essen *et al.*).

621.3.018.41(083.74):621.396.11:551.535 916
Frequency Variations in Short-Wave Propagation—T. Ogawa. (PROC. IRE, vol. 46, pp. 1934–1939; December, 1958.) Frequency measuring equipment, with an accuracy written 1×10^{-8} is described. Observations were made on 5- and 10-mc standard-frequency transmissions from station JJY and the results are discussed in relation ionospheric conditions.

621.317.1.029.6 917
Reciprocity in Radio-Frequency Measurements—G. D. Monteath. (*Electronic Radio Eng.*, vol. 36, pp. 18–20; January, 1959.) A discussion of the advantages of interchangeability of source and detector in measurements of radiation patterns and impedance.

621.317.3:538.632 918
Alternate Current Apparatus for Measuring the Ordinary Hall Coefficient of Ferromagnetic Metals and Semiconductors—J. M. Lavine. (*Rev. Sci. Instr.*, vol. 29, pp. 970–976; November, 1958.) Apparatus for measurements at 1000 c is described having a sensitivity of 10^{-18}w, a noise level of 10^{-9}v and a voltage resolution of 1 in 10^5 for use with sample impedances ranging from less than 1 Ω to several thousand ohms.

621.317.3.018.7:534.78 919
A Sampling Comparator—A. Fischmann-Arbel. (*Electronic Eng.*, vol. 30, pp. 685–689; December, 1958.) The instrument described compares the instantaneous values of two waveforms at constant intervals of time. Applications of the comparator circuit in a delta modulator and a binary quantizer are detailed.

621.317.326 920
Measurement of the Peak Value of a High-Voltage Pulse—G. Giralt. (*Compt. Rend. Acad. Sci., Paris*, vol. 246, pp. 3227–3229; June 9, 1958.) Two methods involving the use of a ballistic galvanometer are described: one in which the rectified current through a standard capacitor is measured [see 1851 of 1957 (Lagasse and Giralt)], the other a blocking method, using a single half-wave rectifier and a circuit-breaking device, applicable for pulses of any waveform.

621.317.335.3.029.6:621.317.733 921
Bridge Method for Microwave Dielectric Measurements—S. H. Glarum. (*Rev. Sci. Instr.*, vol. 29, pp. 1016–1019; November, 1958.) The length of a short-circuited section of liquid-dielectric-filled coaxial line, is adjusted until its input admittance, measured using a microwave bridge, is purely resistive. The dielectric constant, between 1 and 3 kmc obtained by this method, compares well with published values.

621.317.335.3.029.64 922
The Measurement of the Dielectric Properties of Liquids in an H_{01} Resonator—J. S. Dryden. (*J. Sci. Instr.*, vol. 35, pp. 439–440; December, 1958.) A thin-walled silica cup with a flat metal disc at the bottom has proved suitable as a sample container. A method of correction for the meniscus is described.

621.317.34+621.317.38].029.62 923
Triple V.H.F. Reflectometer—G. H. Millard. (*Electronic Radio Eng.*, vol. 36, pp. 11–13; January, 1959.) The instrument is designed for use in the frequency bands 41–68 mc and 88–95 mc for powers of 1 kW or less.

621.317.39:531.76 924
A Catapult End-Speed Recorder—J. R. Pollard. (*Brit. Commun. Electronics*, vol. 5, pp. 676–680; September, 1958.) A description is given of electronic equipment for indicating and recording in printed form the launch speed attained by a steam catapult.

621.317.733:621.314.7 925
A Wide-Band Bridge Yielding Directly the Device Parameters of Junction Transistors—J. Zawels. (IRE Trans. on Electron Devices, vol. ED-5, pp. 21–25; January, 1958. Abstract, Proc. IRE, vol. 46, Part 1, p. 932; May, 1958.)

621.317.75:621.372.413 926
Microwave Double-Sweep Method for Analysis of Time-Dependent Cavity Characteristics—S. Ruthberg. (*Rev. Sci. Instr.*, vol. 29, pp. 999–1003; November, 1958.) A frequency-sweep FM search signal and a frequency-sweep receiver are used to obtain a pulse whose shape is sensitive to cavity resonance. Frequency shift is determined from pulse shape and position along the received signal trace.

621.317.755:621.385.832 927
Oscilloscope Tube with Travelling-Wave Deflection System and Large Field of View—W. F. Niklas and J. Wimpffen. (*J. Brit. IRE*, vol. 18, pp. 653–660; November, 1958.) A tube is described having a balanced helix system as deflection plates and with a large viewing field ($1\frac{3}{8}$ in.$\times$4 in.). Typical operating conditions and performance are described and the relative merits of magnetic and es focusing in a high-speed oscilloscope are discussed in detail.

621.317.799:621.314.7 928
Sweep Equipment Displays Transistor Beta R. Zuleeg and J. Lindmayer. (*Electronics*, vol. 31, pp. 100–101; December 5, 1958.)

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

612.1:621.396.62–519 929
Radio Control of Ventricular Contraction in Experimental Heart Block—M. Verzeano, R. C. Webb, Jr., and M. Kelly. (*Science*, vol. 128, pp. 1003–1005; October 24, 1958.) A method is described for the stimulation of the ventricular myocardium by transmitting a pulse-modulated 2.5-mc carrier which is detected by a receiver enclosed in the chest of the animal under test.

612.84:621.397.5 930
Eye Fixations Recorded on Changing Visual Scenes by the Television Eye-Marker—J. F. Mackworth and N. H. Mackworth. (*J. Opt. Soc. Amer.*, vol. 48, pp. 439–445; July, 1958.) The corneal reflection of a light is picked up by a television camera and is superimposed upon a television monitor displaying the scene which the subject views on a separate screen. The resulting light spot can be made to lie accurately, within one or two degrees, on the part of the scene being regarded by the subject.

621.3.087.5 931
High-Fidelity Video Recording Using Ultrasonic Light Modulation—L. Levi. (*J. Soc. Mot. Pict. Telev. Eng.*, vol. 67, pp. 657–661; October, 1958. Discussion). A method is described for recording video information with bandwidths up to 20 mc on photographic film, using a piezoelectric ultrasonic transducer.

621.3.087.9:621.395.625.3 932
Tones find Data in High-Speed Tape Systems—R. Wasserman and P. Harney. (*Electronics*, vol. 31, pp. 92–95; November 21, 1958.) A digital timing generator, operating during recording, and an associated search unit enable selected data to be extracted from multichannel magnetic-tape systems.

621.365.5:621.316.726.078.3 933
Frequency Stability of R.F. Heating Generators—J. Verstraten. (*Electronic Applic.*, vol. 18, pp. 122–128; August, 1958.) A cavity resonator of large dimensions is used as a tank circuit. Details are given of a 2-kw generator whose frequency remains constant to within 0.5 per cent of 27.12 mc.

621.384.62 934
New Design makes 10-MeV Particle Accelerator Possible—J. L. Danforth. (*Can. Electronics Eng.*, vol. 2, pp. 18–21; July, 1958.) See 935 below.

621.384.62 935
10-MeV Particle Accelerator will Aid Nuclear Research at Chalk River—H. E. Gove. (*Can. Electronics Eng.*, vol. 2, pp. 14–17; July, 1958.) A general description of the tandem Van de Graaff accelerator and associated equipment.

621.385.833 936
Investigations of the Remote-Focus Cathode of Steigerwald—F. W. Braucks. (*Optik, Stuttgart*, vol. 15, pp. 242–260; April, 1958.) Control voltage, and aperture, intensity, position and diameter of the smallest cross section of the beam are discussed in relation to the design parameters of the system.

621.385.833 937
System Design of Asymmetric Unipotential Electron Lenses—K. J. Hanssen. (*Optik, Stuttgart*, vol. 15, pp. 304–317; May, 1958.) See also 1558 of 1957 (Everitt and Hanssen).

621.385.833 938
Shadow-Casting Carbon Films used in Electron Microscopy—A. Oberlin and C. Tchoubar. (*Compt. Rend. Acad. Sci., Paris*, vol. 246, pp. 3329–3332; June 16, 1958.) Carbon films are shown to be preferable to metal for shadow techniques as they have less granulation.

PROPAGATION OF WAVES

621.396.11+621.372.2 939
Transmission and Reflection of Electromagnetic Waves in the Presence of Stratified Media—J. R. Wait. (*J. Res. Nat. Bur. Stand.*, vol. 61, pp. 205–232; September, 1958.) "A general analysis is presented for the electromagmagnetic response of a plane stratified medium consisting of any number of parallel homogeneous layers. The solution is first developed for plane-wave incidence and then generalized to both cylindrical and spherical-wave incidence. Numerical results for interesting special cases are presented and discussed. The application of the results to surface-wave propagation over a stratified ground is considered in some detail."

621.396.11 940
An Example of Guided Propagation in the Mediterranean—L. Boithias and P. Misme. (*Ann. Télécommun.*, vol. 12, pp. 126–132; April, 1957.) Results are analyzed of field-strength measurements at 10 cm λ made on board ship, at distances up to 180 nautical miles from the transmitter, together with meteorological data provided by a tethered radiosonde. Three types of ducting are suggested.

621.396.11:551.510.535 941
Variations in the Direction of Arrival of High-Frequency Radio Waves—J. E. Titheridge. (*J. Atmos. Terr. Phys.*, vol. 13, pp. 17–25; December, 1958.) The effect of propagation in a non-horizontally uniform ionosphere is investigated. The equations are solved for linear and parabolic layers and used to calculate layer tilts necessary to produce the large diurnal bearing changes in the arrival of short waves at Auckland, New Zealand.

621.396.11:551.510.535 942
Very-Long-Distance Ionospheric Propagation—N. C. Gerson. (*J. Atmos. Terr. Phys.*, vol. 13, pp. 169–172; December, 1958.) A brief discussion on the effectiveness of propagation in the spherical shell between the earth and the ionosphere, and also magnetic (whistler-mode) propagation.

621.396.11:551.510.535 943
Long-Distance Single-F-Hop Transmission—W. Dieminger, H. G. Möller and G. Rose. (*J. Atmos. Terr. Phys.*, vol. 13, pp. 191–192; December, 1958.)

621.396.11:551.510.535 944
Magneto-ionic Fading in Pulsed Radio Waves Reflected at Vertical Incidence from the Ionosphere—C. A. Reddy, B. R. Rao, and M. S. Rao. (*J. Brit. IRE*, vol. 18, pp. 669–675; November, 1958.) The difference in phase paths of the two interfering magneto-ionic components is calculated on the basis of ray theory assuming a parabolic electron-density distribution for the F_2 region. The calculated fading frequencies agree fairly well with the observed values. A method of deducing the semi-thickness of the F_2 region from these fading frequencies is described, and some results are given.

621.396.11:551.510.535 945
A New Type of Fading Observable on High-Frequency Radio Transmissions Propagated over Paths Crossing the Magnetic Equator—K. C. Yeh and O. G. Villard, Jr. (Proc. IRE, vol. 46, pp. 1968–1970; December, 1958.) The received energy is split into two independently fading components of comparable strength, separated in frequency by some tens of c. It is

gested that the phenomenon is caused by a nbination of conventional and tilt-supported ›pagation across the evening equatorial ght bulge in the F layer of the ionosphere.

1.396.11:551.510.535:621.3.018.41(083.74) 946
Frequency Variations in Short-Wave Propa-tion—Ogawa. (See 916.)

1.396.11:621.396.812 947
Some Relations between the Bearing and nplitude of a Fading Radio Wave—H. A. hale and L. M. Delves. (*J. Atmos. Terr. ys.*, vol. 13, pp. 72–85; December, 1958.) correlation ratio connecting the changes in nplitude and bearing of a fading wave made › of several randomly phased components is rived. Experimental and theoretical conclu-ons have been compared.

21.396.11.029.45 948
Propagation of Very-Low-Frequency Pulses Great Distances—J. R. Wait. (*J. Res. Nat. ur. Stand.*, vol. 61, pp. 187–203; September, 958.) The space between the earth and the nosphere is represented as a sharply bounded aveguide with concentric spherical boundaries. he concept of phase and group velocity is dis-ussed and applied to determine the influence f the propagation medium on the shape of the nvelope of a quasi-monochromatic pulse. An lternative method is described that is ap-licable to wide-band sources containing many pectral components.

RECEPTION

21.372.632.029.6 949
Design Considerations in a Wide-Band Microwave Mixer and I. F. Preamplifier—J. C. Rennie. (IRE TRANS. ON COMMUNICATIONS SYSTEMS vol. CS-5, pp. 21–25; September, 957.)

521.396.621.22:621.375.121.2 950
An Electronic Multicoupler and Antenna Amplifier for the V.H.F. Range—K. Fischer. (IRE TRANS. ON COMMUNICATIONS SYSTEMS, vol. CS-5, pp. 43–48; December, 1957. Abstract, PROC. IRE, vol. 46, p. 515; February, 1958.)

621.396.66:621.316.726.078.3 951
A.F.C. in Band-II F.M. Receivers: using a Junction Diode—G. D. Browne. (*Mullard Tech. Commun.*, vol. 4, pp. 152–157; November, 1958.) The reactance variation obtained by varying the Ge-diode reverse voltage provides a frequency control operating successfully above 10 mc.

621.396.666 952
Laboratory Tests on Kahn's Theory of Anti-fading Reception—G. Bronzi. (*Alta Frequenza*, vol. 27, pp. 17–43; February, 1958.) Analysis of an experimental comparison of methods of double-diversity reception using strongest-signal selection or the ratio-squarer combining system [(541 of 1955) Kahn] shows the advantages of the latter method.

621.396.81 953
Correlation Measurements in the Short-Wave Range—J. Grosskopf, M. Scholz, and K. Vogt. (*Nachrichtentech. Z.*, vol. 11, pp. 91–95; February, 1958.) Report on measurements of the correlation coefficient in diversity reception using rhombic and dipole antennas.

621.396.82 954
Radio Interference: Part 5—Industrial, Scientific and Medical Apparatus and Radiating Receivers—C. W. Sowton and A. C. R. Britton. (*P.O. Elect. Engr. J.*, vol. 51, Part 3, pp. 202–205; October, 1958.) International frequency allocations for industrial, scientific, and medical apparatus are discussed and examples of the field strength of radiation at fundamental and harmonic frequencies are tabulated. The free-radiation frequency of 27.12 mc with harmonic attenuation has been adopted for medical apparatus. Methods for measuring the radiation field receivers are outlined. These methods have been accepted by most countries. Part 4: 3964 of 1958 (Macpherson).

621.396.822:621.397.62 955
Design Considerations in the Reduction of Sweep Interference from Television Receivers —A. M. Intrator. (IRE TRANS. ON BROADCAST AND TELEVISION RECEIVERS, vol. BTR-2, pp. 1–5; April, 1956.)

STATIONS AND COMMUNICATION SYSTEMS

621.376.5 956
Using Markerless Pulse Trains to Communicate—M. Davidson, H. Joseph, and N. Zucker. (*Electronics*, vol. 31, pp. 89–91; November 21, 1958.) Three types of markerless pulse-train modulation are compared, and demodulating circuits for pulse-interval modulation are described.

621.376.5:621.391 957
Statistics of Regenerative Digital Transmission—W. R. Bennett. (*Bell Syst. Tech. J.*, vol. 37, pp. 1501–1542; November, 1958.) Specific problems considered include the properties of a digital message pulse train as a random noise source, the effect of time jitter in a received pulse train on the recovered analogue signal, and the derivation of the pulse repetition frequency from a pulse train by shock excitation of a tuned circuit.

621.391 958
The Central Concepts of Communication Theory for Infinite Alphabets—I. Fleischer. (*J. Math. Phys.*, vol. 37, pp. 223–228; October, 1958.)

621.391:621.376 959
Demodulation and Detection—D. A. Bell. (*Electronic Radio Eng.*, vol. 36, pp. 21–24; January, 1959.) A discussion of the basic processes, particularly in relation to information theory.

621.394.3:621.396.65 960
Teleprinting over Long-Distance Radio Links—A. C. Croisdale. (*P.O. Elec. Eng. J.*, vol. 51, Parts 2 and 3, pp. 88–93 and 219–225; July and October, 1958.) Methods for the transmission of 5-unit telegraphy signals and techniques for reducing the mutilation of signals on radio links are described. Automatic error correction and the resulting improvement in performance are discussed with reference to CCITT recommendations.

621396(667) 961
Radio Communications in Ghana—R. G. Sharpe and D. R. Gamlen. (*Brit. Commun. Electronics*, vol. 5, pp. 750–756; October, 1958.)

621.396.2 962
The Bandwidth Occupied by a Class A1 Transmission and its Determination—J. Marique. (*Rev. HF, Brussels*, vol. 3, no. 10, pp. 359–368; 1957.)

621.396.2 963
Performance of Some Radio Systems in the Presence of Thermal and Atmospheric Noise—A. D. Watt, R. M. Coon, E. L. Maxwell, and R. W. Plush. (PROC. IRE, vol. 46, pp. 1914–1923; December, 1958.) "The performance of several basic types of communication systems are determined experimentally, and in some cases theoretically, under typical conditions with steady or fading carriers, and in the presence of thermal or atmospheric noise. The relative efficiency of various carriers and the interference factor of various types of noise are found to be dependent upon the characteristics of the particular communication system as well as the characteristics of the carrier and noise themselves. Methods are considered for calculating errors expected from a given system, based upon the amplitude distribution of the noise envelope."

621.396.2:[629.19+523.3 964
Long-Distance Telecommunications by means of Satellites—F. Vilbig. (*Elektrotech. Z.*, vol. 79, pp. 375–382; June 1, 1958.) Propagation conditions and operational problems relating to the use of the moon and of artificial satellites as reflectors or relay stations are discussed.

621.396.25:621.394 965
High-Speed Frequency-Shift Keying of L.F. and V.L.F. Radio Circuits—H. G. Wolff. (IRE TRANS. ON COMMUNICATIONS SYSTEMS, vol. CS-5, pp. 29–42; December, 1957. Abstract, PROC. IRE, vol. 46, p. 515; February, 1958.)

621.396.3 966
Automatic Error Correction—P. R. Keller and L. K. Wheeler. (*Wireless World*, vol. 65, pp. 28–33; January, 1959.) A description, with block diagrams, of the "Autoplex" two-channel time-division electronic equipment, and of the improvement in the error rate with automatic repetition. See also 2224 of 1958 (Keller) and 960 above.

621.396.4:621.376.5 967
The Timing of High-Speed Regenerative Repeaters—O. E. De Lange. (*Bell Syst. Tech. J.*, vol. 37, pp. 1455–1486; November, 1958.) A simplified method for determining the performance of the timing portion of chains of regenerative repeaters in multi-channel PCM systems. Two types are considered, those in which a repetitive timing signal is sent on one channel, and those in which timing is obtained from the coded pulse trains. The effect of random noise is calculated and other transmission defects are discussed.

621.396.4:621.376.5 968
Experiments on the Timing of Regenerative Repeaters—O. E. De Lange and M. Pustelnyk. (*Bell Sys. Tech. J.*, vol. 37, pp. 1487–1500; November, 1958.) Experiments performed with self-timed binary regenerative repeaters to determine the behavior of the timing portion of a chain of such repeaters are described. The number of errors produced by the action of noise on the timing system was negligible compared with those produced by other effects of noise.

621.396.4:621.376.5 969
Timing in a Long Chain of Regenerative Binary Repeaters—H. E. Rowe. (*Bell Syst. Tech. J.*, vol. 37, pp. 1543–1598; November, 1958.) The power spectra and the total powers of the timing noise, spacing noise, and alignment noise, caused by the input noise at each repeater are determined for a long chain of regenerative repeaters using either tuned-circuit or locked-oscillator timing filters. Tuning error effects are studied for a repeater chain using locked-oscillator timing circuits.

621.396.41 970
The 'Third Method'—J. F. H. Aspinwall. (*Wireless World*, vol. 65, pp. 39–43; January, 1959.) The method described by D. K. Weaver, Jr. (PROC. IRE, vol. 44, pp. 1703–1705; December, 1956.) is compared with filter and phasing methods of generating ssb signals, and its application to radio telephony is briefly described.

621.396.41 971
A Compatible Single-Sideband Modulation System—L. R. Kahn. (*Proc. Radio Club. Amer.*, vol. 34, pp. 3–9; March, 1958.) A system is described which is compatible with the existing dsb am system. It also compares favorably in signal/noise ratio, spectrum economy and reduction of selective fading distortion with the conventional ssb system, and it is suitable for aeronautical communications. Operational tests show that the greatest improvement in audio fidelity and signal/noise ratio occurs in narrow-bandwidth domestic receivers. See 3247 of 1958 (Costas) for mathematical analysis of the system.

621.396.41:551.510.52 972
Quadruple-Diversity Tropospheric Scatter Systems—W. G. Long and R. R. Weeks. (IRE TRANS. ON COMMUNICATIONS SYSTEMS, vol. CS-5, pp. 8–19; December, 1957.)

621.396.41:621.396.65 973
Microwave Network now Spans Canada—M. James. (*Can. Electronics Eng.*, vol. 2, pp. 28–31; September, 1958.) A general description of the TD-2 multiplex FM system. See also 3638 of 1958 (Curtis et al.).

621.396.5 974
Practical and Theoretical Design Considerations for Bridge Negative-Feedback Amplifiers in Carrier Telephony—M. J. Cotterill and J. W. Halina. (IRE TRANS. ON COMMUNICATIONS SYSTEMS, vol. CS-5, pp. 26–31; September, 1957.)

621.396.65 975
Scatter Equipment Built for Canadian Use—J. A. Grant. (*Can. Electronics Eng.*, vol. 2, pp. 30–33; July, 1958.) Terminal and repeater FM equipment covering the range 755–980 mc is described, with particular reference to the "serrasoid" sawtooth modulator.

621.396.65:621.396.967 976
Microwave Links for Radar Networks—J. W. Sutherland. (*Brit. Commun. Electronics*, vol. 5, pp. 688–695; September, 1958.) The operational advantages of remote presentation of radar information and the parameters of practical transmission systems are discussed in detail. The use of travelling-wave tubes in IF and nondemodulating types of repeater is described.

621.396.7:621.396.11 977
Radio Propagation Transmitting Station WWI at Havana, Illinois—(*Tech. News Bull. Nat. Bur. Stand.*, vol. 42, pp. 154–155; August, 1958.) The station, which is centered at 40° 13.27′ N, 90° 1.39′ W, is intended mainly for research in VHF transmission. Its twelve transmitters operate on 30–108 mc at 3–50 kw. Details of the antenna system are given.

SUBSIDIARY APPARATUS

621.311.62:621.314.7 978
Boosting Power-Transistor Efficiency—J. W. Caldwell and T. C. G. Wagner. (*Electronics*, vol. 31, pp. 86–88; November 21, 1958.) High efficiency is obtained by carefully controlling the instantaneous voltage and current through the transistor.

621.316.721:621.318.3:621.314.7 979
Precision Current Regulator using Transistors—S. D. Johnson and J. R. Singer. (*Rev. Sci. Instr.*, vol. 29, pp. 1026–1028; November, 1958.) A regulator for electromagnets of field strength between 1000 and 5000 G is described which is accurate to within two parts in 10^5.

621.316.722:621.314.7 980
Transistor Voltage Regulators—G. W. Meszaros. (*Bell Lab. Rec.*, vol. 36, pp. 442–445; December, 1958.) Design considerations are discussed for series-type regulators and the magnetic-amplifier type designed for the TRIDAC computer is described.

TELEVISION AND PHOTOTELEGRAPHY

621.397.24:621.3.018.782.4 981
Demonstration of Delay Distortion Correction by Time-Reversal Techniques—B. P. Bogert. (IRE TRANS. ON COMMUNICATIONS SYSTEMS vol. CS-5, pp. 2–7; December, 1957. Abstract, PROC. IRE, vol. 46, p. 515; February, 1958.)

621.397.5 982
The Possibilities of Reduced Television Bandwidth—S. Deutsch. (IRE TRANS. ON BROADCAST AND TELEVISION RECEIVERS, vol. BTR-2, pp. 69–82; October, 1956. Abstract, PROC. IRE, vol. 45, p. 253; February, 1957.)

621.397.5:621.396.4 983
Study of Multichannel Sound Transmission from a Single Transmitter: Application to Bilingual Television—L. Bourassin. (*Électronique, Paris*, pp. 45–50 and 37–44; May and July/August, 1957.) Continuation of 3312 of 1957.

621.397.611.001.4 984
Two New B.B.C. Transparencies for Testing Television Camera Channels—(*BBC Eng. Div. Monographs*, n.p. pp. 5–17; November, 1958.) **[Part 1—Requirements, Design and Use of the Colour Response, and the Gradation and Resolution Transparencies**—G. Hersee (pp. 5-13). **[Part 2—The Manufacture of the B.B.C. Test Transparency No. 51**—J. R. T. Royle (pp. 14–17).

621.397.611.2 985
The Problem of Inertia Effects in Television Camera Tubes of the Vidicon Type—C. Kunze. (*Hochfreq. und Elektroak.*, vol. 66, pp. 84–89; November, 1957.) Measurements show that inertia effects are mainly due to incomplete recharging of the picture elements [see also 252 of 1956 (Heimann)]. The dependence of these effects on operating parameters, and methods of eliminating inertia effects, such as roughening the target surface, are discussed.

621.397.62:535.623 986
Tentative Methods of Measurement of Colour Television Receiver Performance—S. P. Ronzheimer and R. J. Farber. (IRE TRANS. ON BROADCAST AND TELEVISION RECEIVERS, vol. BTR-2, pp. 10–30; April, 1956.)

621.397.62:621.314.7 987
Transistors in Television Receivers—B. R. Overton. (*J. Telev. Soc.*, vol. 8, pp. 444–468; July–September, 1958.) "A complete receiver, operating from a 12-v battery (consumption 12 w approximately), and employing transistors throughout is described. Special attention is given to the three major technical problems of incorporating transistors in television receivers, namely, RF and IF amplification, video drive and line scanning. The paper discusses some new techniques for dealing with these problems, in particular a system of scan magnification, hitherto undisclosed. Some attempt is made to forecast future trends."

621.397.62:621.372.54 988
Trap Improves TV Picture—G. C. Field. (*Electronics*, vol. 31, pp. 100, 102; November 21, 1958.) The use of a bifilar-T trap to suppress adjacent-channel interference in 40-mc IF amplifiers is described.

621.397.621.2 989
Improvements in Television Receivers: Part 4—Stabilization of the Line Deflection Circuit by means of a V.D.R. Resistor—B. G. Dammers, A. G. W. Uitjens, A. Boekhorst, and H. Heyligers. (*Electronic Applic.*, vol. 18, pp. 118–121; August, 1958.) A simplified version of the circuit proposed in 3316 of 1957 makes the protection device proposed in 282 of January (Dammers *et al.*) unnecessary. A voltage-dependent resistor is used to control the line output valve. The scanning current is stabilized to about ±2 per cent for line voltage variations of ±13 per cent.

621.397.2:535.623:621.385.832 990
Error Correction in Mask-Type Colour Television Tubes—S. H. Kaplan. (*J. Telev. Soc.*, vol. 8, pp. 470–480; July–September, 1958.) Corrections for the errors which increase as a function of scan angle, and optical exposure methods for correcting triad sire and location errors, are described. A proposal to eliminate triad shape errors by means of a radially distorted aperture mask pattern is given in detail.

621.397.7 991
B.B.C. Install 'Translator' for Improved TV Reception—(*Brit. Commun. Electronics*, vol. 5, p. 757; October, 1958.) Signals transmitted on one channel are converted, without demodulation, to another and automatically reradiated at low power over line-of-sight paths to the area to be served. Equipment at Folkestone is briefly described.

621.397.7:621.315.212 992
Interconnection of Television Cable Links at the Carrier-Frequency State according to the C.C.I.F.—R. Hoffmann. (*Nachrichtentech. Z.*, vol. 11, pp. 96–99; February, 1958.) Summary of the CCIF recommendations of December, 1956, concerning the transmission of television signals over international cable routes.

621.397.8:535.623 993
The Origin and Measurement of Level-Dependent Phase and Amplitude Fluctuations in the Transmission of the Subcarrier in Colour Television—J. Piening. (*Nachrichtentech. Z.*, vol. 11, pp. 70–77; February, 1958.) Causes of amplitude and phase distortion are investigated quantitatively.

TUBES AND THERMIONICS

621.314.63:546.289 994
New Types of Germanium Diodes and their Circuit Applications—G. Grimsdell. (*Electronic Eng.*, vol. 30, pp. 709–710; December, 1958.) The properties of gold-bonded diodes and small-area junction diodes are compared with those of point-contact types and the relative advantages and disadvantages are outlined.

621.314.7 995
On the Variation of Transistor Small-Signal Parameters with Emitter Current and Collector Voltage—N. I. Meyer. (*J. Electronics Control*, vol. 5, pp. 329–337; October, 1958.) Addendum to 3285 of 1958.

621.314.7 996
The Characteristic Frequencies of a Junction Transistor—J. M. Rollett. (*J. Electronics Control*, vol. 5, pp. 344–347; October, 1958.) The relations between characteristic frequencies and intrinsic parameters are considered using the model which assumes one-dimensional flow of carriers across the base region. The frequencies discussed are the cut-off frequencies in the common-base and common-emitter connections, and the frequency at which the common-emitter current gain is unity.

621.314.7 997
Structure-Determined Gain-Band Product of Junction-Triode Transistors—J. M. Early. (PROC. IRE, vol. 46, pp. 1924–1927; December, 1958.) The fundamental frequency limitations

e discussed, and it is shown that mesa-type ansistors for use in the microwave region are heoretically possible. The gain-bandwidth roduct is an order of magnitude better than hat of field-effect or analogue transistors.

21.314.7 998
Theory of the *P-N* Junction Device using valanche Multiplication—T. Misawa. (PROC. RE, vol. 46, p. 1954; December, 1958.)

21.314.7 999
The Internal Current Gain of Drift Transistors—F. J. Hyde. (PROC. IRE, vol. 46, pp. 963–1964; December, 1958.)

21.314.7:546.289 1000
High-Frequency Germanium Transistors—. S. Lamming. (*Research, London*, vol. 11, pp. 25–431; November, 1958.) Design theory and roduction techniques are reviewed.

21.314.7:546.289 1001
Very-High-Power Transistors with Evaporated Aluminium Electrodes—H. W. Henkels nd G. Strull. (IRE TRANS. ON ELECTRON DEVICES, vol. ED-4, pp. 291–294; October, 1957. Abstract, PROC. IRE, vol. 46, p. 515; February, 1958.)

621.314.7:621.317.733 1002
A Wide-Band Bridge Yielding Directly the Device Parameters of Junction Transistors—J. Zawels. (IRE TRANS. ON ELECTRON DEVICES, vol. ED-5, pp. 21–25; January, 1958. Abstract, PROC. IRE, vol. 46, Part 1, p. 932; May, 1958.)

621.314.7:621.317.799 1003
Sweep Equipment Displays Transistor Beta—R. Zuleeg and J. Lindmayer. (*Electronics*, vol. 31, pp. 100–101; December 5, 1958.)

621.314.7:621.318.57 1004
A New High-Current Mode of Transistor Operation—C. G. Thornton, and C. D. Simmons. (IRE TRANS. ON ELECTRON DEVICES, vol. ED-5, pp. 6–10; January, 1958. Abstract, PROC. IRE, vol. 46, Part 1, p. 932; May, 1958.)

611.314.7:621.318.57 1005
The 'Thyristor'—a New High-Speed Switching Transistor—C. W. Mueller and J. Hilibrand. (IRE TRANS. ON ELECTRON DEVICES, vol. ED-5, pp. 2–5; January, 1958. Abstract, PROC. IRE, vol. 46, Part 1, pp. 931–932; May, 1958.)

621.314.7:621.318.57 1006
A Transistor with Thyratron Characteristics and Related Devices—W. von Münch. (*J. Brit. IRE*, vol. 18, pp. 645–652; November, 1958.) Details of production and electrical performance are given for a device produced by immersing a tungsten whisker in the collector of an n-p-n-junction transistor, of high base resistivity, during the alloy process. Devices with more than one output electrode are developed from this design. A special structure for triggering by radiation and a symmetrical switching transistor are also studied.

621.314.7.002.2:546.28 1007
Silicon Transistors—J. T. Kendall. (*Research, London*, vol. 11, pp. 381–386; October, 1958.) Industrial manufacturing techniques are described.

621.314.7.01 1008
On the Need for Revision in Transistor Terminology and Notation—H. L. Armstrong. (PROC. IRE vol. 46, pp. 1949–1950; December, 1958.)

621.314.7.012.8 1009
Transistor Equivalent Circuits—C. Moerder. (*Elektrotech. Z.*, vol. 79, pp. 469–472; July 1, 1958.) Derivation of equivalent circuits from fundamental considerations.

621.383.4 1010
Properties of Cadmium Sulphide Photoconductive Cells—R. L. Williams. (*Can. J. Phys.*, vol. 36, pp. 1536–1550; November, 1958.)

621.383.49 1011
The Characteristics of Evaporated CdS and CdSe Photistors—D. A. Anderson. (*Can. Electronics Eng.*, vol. 2, pp. 23–29; July, 1958.) The performance of various types of photoconductive cell is discussed.

621.383.8:535.215–15 1012
High-Sensitivity Crystal Infrared Detectors—M. E. Lasser, P. Cholet, and E. C. Wurst, Jr. (*J. Opt. Soc. Amer.*, vol. 48, pp. 468–473; July, 1958.) The characteristics of n-type Au-doped Ge photoconductive cells are tabulated, and comparisons are made with the photovoltaic p-n-junction InSb cell. A multiple-contact cell is described which can locate a target as well as detect it without a moving optical system [see also 3691 of 1957 (Wallmark)].

621.385:534.39 1013
Analysis of Microphony in Electron Tubes—A. Stecker. (*Electronic Applic.*, vol. 18, pp. 99–117; August, 1958.)

621.385.029.6 1014
On the Coupling Coefficients in the 'Coupled-Mode' Theory—A. Yariv. (PROC. IRE, vol. 46, pp. 1956–1957; December, 1958.) Perturbation theory is used to evaluate the coupling coefficients for the case of small coupling, taking the field and current values as those applicable to the no-coupling case.

621.385.029.6 1015
Effect of Beam Coupling Coefficient on Broad-Band Operation of Multicavity Klystrons—S. V. Yadavalli. (PROC. IRE, vol. 46, pp. 1957–1958; December, 1958.) Wide-band multicavity klystrons are more efficient in the L band of frequencies than in the S band, and are better operated at higher voltages than synchronously tuned klystrons.

621.385.029.6 1016
Pulse-Modulated Beam Current Improves Operation of Mixer-Series Klystrons—A. K. Scrivens. (*Can. Electronics Eng.*, vol. 2, pp. 32–34; October, 1958.)

621.385.029.6 1017
On the Design of the Transition Region of Axisymmetric, Magnetically Focused Beam Valves—V. Bevc, J. L. Palmer, and C. Süsskind. (*J. Brit. IRE*, vol. 18, pp. 696–708; December, 1958.) A method is described for using an analogue computer to trace out electron trajectories. It is applied to the presentation of beam envelopes for Brillouin flow, periodic magnetic focusing and space-charge-balanced flow. By matching these with envelopes derived from the theory of the Pierce gun, it is possible to specify the dimensions of a gun for producing a required beam.

621.385.029.6 1018
A Relativistic Treatment of Space-Charge-Limited Current in a Planar Diode Magnetron before Cut-Off—J. A. Bradshaw. (*J. Electronics Control*, vol. 5, pp. 300–306; October, 1958.) "Relativistic expressions, correct to first order in a current density parameter, are obtained from two cross-coupled integral equations for the potential functions in a planar diode magnetron. These expressions are evaluated and plotted in two figures, and compared with expressions given by Gold [3695 of 1957]."

621.385.029.6 1019
Travelling-Wave Valves—J. Voge. (*Ann. Télécommun.*, vol. 12, pp. 92–104 and 105–119; March and April, 1957.) A general survey of the development of the device. Carcinotrons and amplifiers are described and comparative tables of French and American types are given. Amplifier noise and methods for reducing it, and the characteristics of several linear accelerators are discussed. 28 references.

621.385.029.6 1020
Wave Matrices Applied to a Periodically Loaded Travelling-Wave Tube—D. E. T. F. Ashby. (*J. Electronics Control*, vol. 5, pp. 338–343; October, 1958.) The derivation of the matrices is shown, and they are used to determine the conditions necessary for oscillation due to feedback caused by reflections from the periodic discontinuities. Oscillations may be produced unintentionally, and be mistaken for backward-wave oscillations.

621.385.029.6:621.318.2 1021
The Design of Periodic Magnetic Focusing Structures—J. E. Sterrett and H. Heffner. (IRE TRANS. ON ELECTRON DEVICES, vol. ED-5, pp. 35–42; January, 1958. Abstract, PROC. IRE, vol. 46, Part 1, p. 932; May, 1958.)

621.385.029.6:621.372.2 1022
A Note on the Dispersion of Interdigital Delay Lines—F. Paschke. (*RCA Rev.*, vol. 19, pp. 418–422; September, 1958.) "It is shown that the effect of the backwall on the dispersion of an interdigital delay line can be taken into account by a lumped susceptance which periodically loads the 'ideal' line. Experimental results are in good agreement with the theory." See also 629 of 1957.

621.385.029.6:621.396.822 1023
Noise Wave Excitation at the Cathode of a Microwave Beam Amplifier—W. R. Beam. (IRE TRANS. ON ELECTRON DEVICES, vol. ED-4, pp. 226–234; July, 1957. Abstract, PROC. IRE, vol. 45, p. 1760; See also 4067 of 1957 (Knechtli and Beam).

621.385.029.621.63 1024
Travelling-Wave Amplifiers and Backward-Wave Oscillators for V.H.F.—D. A. Dunn. (IRE TRANS. ON ELECTRON DEVICES, vol. ED-4, pp. 246–264; July, 1957. Abstract, PROC. IRE, vol. 45, pp. 1760–1761; December, 1957.)

621.385.029.64:537.533:621.375.9 1025
Parametric Amplification of Space-Charge Waves—A. Ashkin. (*J. Appl. Phys.*, vol. 29, pp. 1646–1651; December, 1958.) An experimental investigation of the theory of Louisell and Quate (2273 of 1958), which shows that a signal imposed on a beam as a "fast" or "slow" space-charge wave can be exponentially amplified by strongly modulating the beam with a RF wave of twice the signal frequency. Over a 10-inch length of beam a 41-db increase has been observed. With the high-level signal frequency lower than the signal frequency, an increase of 30 db over a 9.2-inch length was observed.

621.385.032.213.13 1026
On the Conduction Mechanism of Oxide-Coated Cathode—H. Mizuno. (*J. Phys. Soc. Japan*, vol. 13, pp. 1234–1235; October, 1958.)

621.385.032.213.13:621.396.822 1027
Noise in Oxide Cathode Coatings—H. J. Hannam and A. van der Ziel. (*J. Appl. Phys.*, vol. 29, pp. 1702–1705; December, 1958.) A discussion of noise measurements at 8 mc and 30 c. The HF measurements show thermal noise at high and low cathode temperatures with a pronounced noise peak caused by shot noise where pore conduction changes to grain conduction. At LF the results show that the pores are inherently noisier than the grains.

621.385.032.213.13:621.396.822 1028
New Mechanism for the Generation of Flicker Noise—F. Fisher and I. P. Valkó. (*J. Appl. Phys.*, vol. 29, p. 1772; December, 1958.) The effect of cathode porosity on flicker noise in valves has been investigated, and results confirm the theoretical predictions of Lindemann and van der Ziel (1284 of 1957).

621.385.032.213.63 1029
The Breakdown of Cathode Coatings—B. Wolk. (*Sylvania Technologist*, vol. 10, pp. 106–110; October, 1957.) A report is given of an experimental study of the processing of cathode coatings under a set of controlled temperature and time conditions. Results indicate that thermal decomposition characteristics and particle size are related. Factors accounting for poor processing reliability are also considered.

621.385.032.24 1030
The Grid Emitting Properties of Titanium—J. A. Champion. (*Brit. J. Appl. Phys.*, J. A. Champion. (*Brit. J. Appl. Phys.*, vol. 9, pp. 491–495; December, 1958.) Experiments show that titanium is suitable as a screen-grid winding wire and for other electrode applications when used in the temperature range 700–900°C; above this cathode poisoning occurs.

621.385.032.26 1031
Dynamics of Electron Beams from Magnetically Shielded Guns—A. Ashkin. (*J. Appl. Phys.*, vol. 29, pp. 1594–1604; November, 1958.) Theoretical and experimental investigation of electron orbits and beam shapes. Current density variations and transverse velocity are measured with a beam analyser. In the limit of negligible space charge the observations are explained on the basis of simple energy considerations, taking account of the effect of transverse thermal velocities. In the presence of space charge the beam behavior is qualitatively unchanged, provided that the magnetic field is higher than about three times the Brillouin field.

621.385.3 1032
New Electron Tubes for Wide-Band Amplifiers—S. Edsman. (*Ericsson Rev.*, vol. 35, no. 3, pp. 98–102; 1958.) A note on the characteristics and operation of a pentode Type 5847/404A, triode Type 5842/417A, and tetrode Type 7150.

621.385.3:621.365.5 1033
Two Methods of Calculation for the Class-C Operation of Transmitter Valves—N. Weyss. (*Elektrotech. u. Maschinenb.*, vol. 75, pp. 633–638; December 1, 1958.) Oscillators for RF heating are considered.

621.385.832:621.317.755 1034
Oscilloscope Tube with Travelling-Wave Deflection System and Large Field of View—Niklas and Wimpffen. (See 927.)

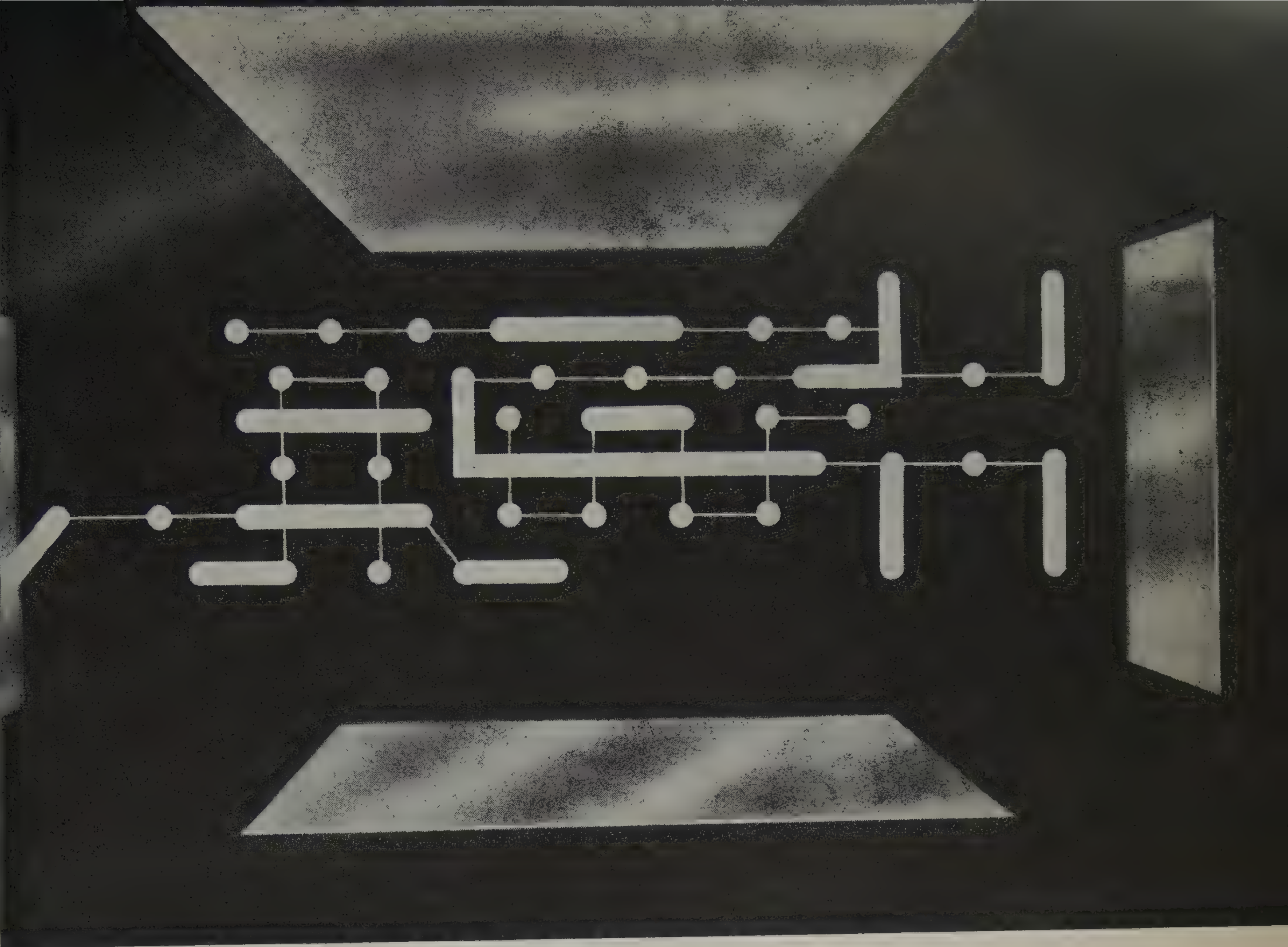
621.385.832:621.397.621.2:666.1 1035
A Method of Sealing the Window and Cone of Television Picture Tubes—A. H. Edens. (*Philips Tech. Rev.*, vol. 19, pp. 318–323; May 31, 1958. *Glass Ind.*, vol. 39, pp. 534–538; October, 1958.)

MISCELLANEOUS

061.6:621.396 1036
International Radio Organizations—some Aspects of their Work—R. L. Smith-Rose. (*J. Brit. IRE*, vol. 18, pp. 631–639; November, 1958.) An historical outline is given of the growth of communications and the resulting organizations set up to control international affairs. Details of the present work of the International Radio Consultative Committee (CCIR) are given and the activities of the International Scientific Radio Union (URSI) are reviewed.

621.3.002.5 1037
Production Machinery for the Electronics Industry—G. Sideris. (*Electronics*, vol. 31, pp. 73–84; October 24, 1958.) A review dealing with the modernization of machinery, tools, and plant layout, and with advances in component production and equipment assembly techniques.

621.37/.38].004.1 1038
The Influence of Interaction Reliabilities—M. A. Acheson. (*Sylvania Technologist*, vol. 11, pp. 91–95; July, 1958.) Examples are given of simple electronic systems in which interaction between two or more reliabilities of parts influences the total reliability.



Report from IBM ▲ *Yorktown Research Center, New York*

LOW TEMPERATURES FOR HIGH-SPEED CIRCUITRY

Certain metals and alloys lose their resistance to electricity at temperatures close to absolute zero. They become "super-conductors." Investigations by Dr. D. R. Young and others at the IBM Yorktown Research Center are directed toward the utilization of this unique property in the development of small-size, high-speed switches with increased logical capacities.

Interestingly, when a "super-conductor" is exposed to certain magnetic fields, it reverts instantly to an ordinary conductor. One experimental switching device that takes advantage of this property has been constructed at IBM. In essence, it is a "sandwich" of glass, tin, silicon monoxide and lead. The device is immersed in liquid helium to bring it close to absolute zero. The tin strip becomes a "super-conductor," so current flows readily through it. When a current is applied to the lead strip it creates a magnetic field. As a result, the tin strip is no longer a "super-conductor" and now has electrical resistance . . . it is "off." Remove the magnetic field and it is "on" again. This then is an "on-off" device, or switch, that is expected to work at speeds much greater than present switch capacities. There are no moving parts to wear out and 1,000 such devices can be mounted on a bit of glass only a few inches square.

In addition to these experiments, the study of matter at very low temperatures is being applied to other areas at IBM. The immediate objective is to apply the results to the development of device formulations which will greatly accelerate arithmetic speed and increase the logical capacity in electronic computers of greatly reduced size.

IBM® RESEARCH

Investigate the many career opportunities available in exciting new fields at IBM.

DELCO POWER TRANSISTORS

NEW TYPES

MILITARY
COMMERCIAL

TYPICAL CHARACTERISTICS AT 25°C

EIA	2N297A*	2N297A	2N665**	2N553
Collector Diode Voltage (Max.)	60	60	80	80 volts
HFE ($I_c = 0.5A$) (Range)	40-100	40-100	40-80	40-80
HFE ($I_c = 2A$) (Min.)	20	20	20	20
I_{co} (2 volts, 25°C) (Max.)	200	200	50	50 μa
I_{co} (30 volts, 71°C) (Max.)	6	6	2	2 ma
F_{ae} (Min.)	5	5	20	20 kc
T (Max.)	95	95	95	95°C
Therm Res. (Max.)	2	2	2	2° C/W

*Mil. T 19500/36 (Sig. C.)
**Mil. T 19500/58 (Sig. C.)

NOTE: Military Types pass comprehensive electrical tests with a combined acceptance level of 1%.

Delco Radio announces new PNP germanium transistors in 2N553 series—the 2N297A and 2N665, designed to meet military specifications. These transistors are ideal as voltage and current regulators because of their extremely low leakage current characteristics. All are highly efficient in switching circuits and in servo amplifier applications, and all are in *volume* production! Write today for complete engineering data.

DELCO RADIO

Division of General Motors • Kokomo, Indiana

BRANCH OFFICES

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726 Santa Monica Boulevard
Tel: Exbrook 3-1465

Professional Group Meetings

(*Continued from page 90A*)

Detroit—December 3

"The Datamatic 1000 Computer," G. W. Pratt, E. A. Schroeder, and A. Carr, Datamatic Corp.

Philadelphia—January 8

"Automatic Correlation of Data in the Differential Diagnosis of Hematologic Diseases," R. A. C. Lane, RCA.

Philadelphia—February 5

"Matrix Logic," E. J. Schubert, Burroughs Corp.

Pittsburgh—October 22

"Gamma 60 Computer," W. Taransky, Sperry-Rand Corp.

"Datamatic-1000 Computer," C. L. Dudley, Jr., Minneapolis-Honeywell Regulator Co.

Pittsburgh—November 12

"Design of a Very High-Speed Computer," D. B. Gillies, Univ. of Illinois.

Pittsburgh—November 18

"Optimizing Control Systems for the Process Industries," D. A. Burt, Westinghouse Electric Corp.

San Francisco—January 20

"Redundancy Techniques for Computer Reliability," L. Fein.

Washington, D. C.—February 11

"Magnetic Ink Character Reading," G. Miller, Gen. Elec. Co.

Engineering Management

Boston—December 4

"Performance Evaluation from the Management Point of View," D. C. Blumer, Natl. Metal Trades Assoc.

Boston—January 29

"Raytheon's Approach to Engineering Management Selection and Development," A. C. Hilton, Raytheon Mfg. Co.

Philadelphia—January 22

"Competitive Fixed-Price Creativity—Industrial Research," L. Steg, Gen. Elec. Co.

"Competitive Fixed-Price Creativity—Government Research," A. C. Munster, Philco Corp.

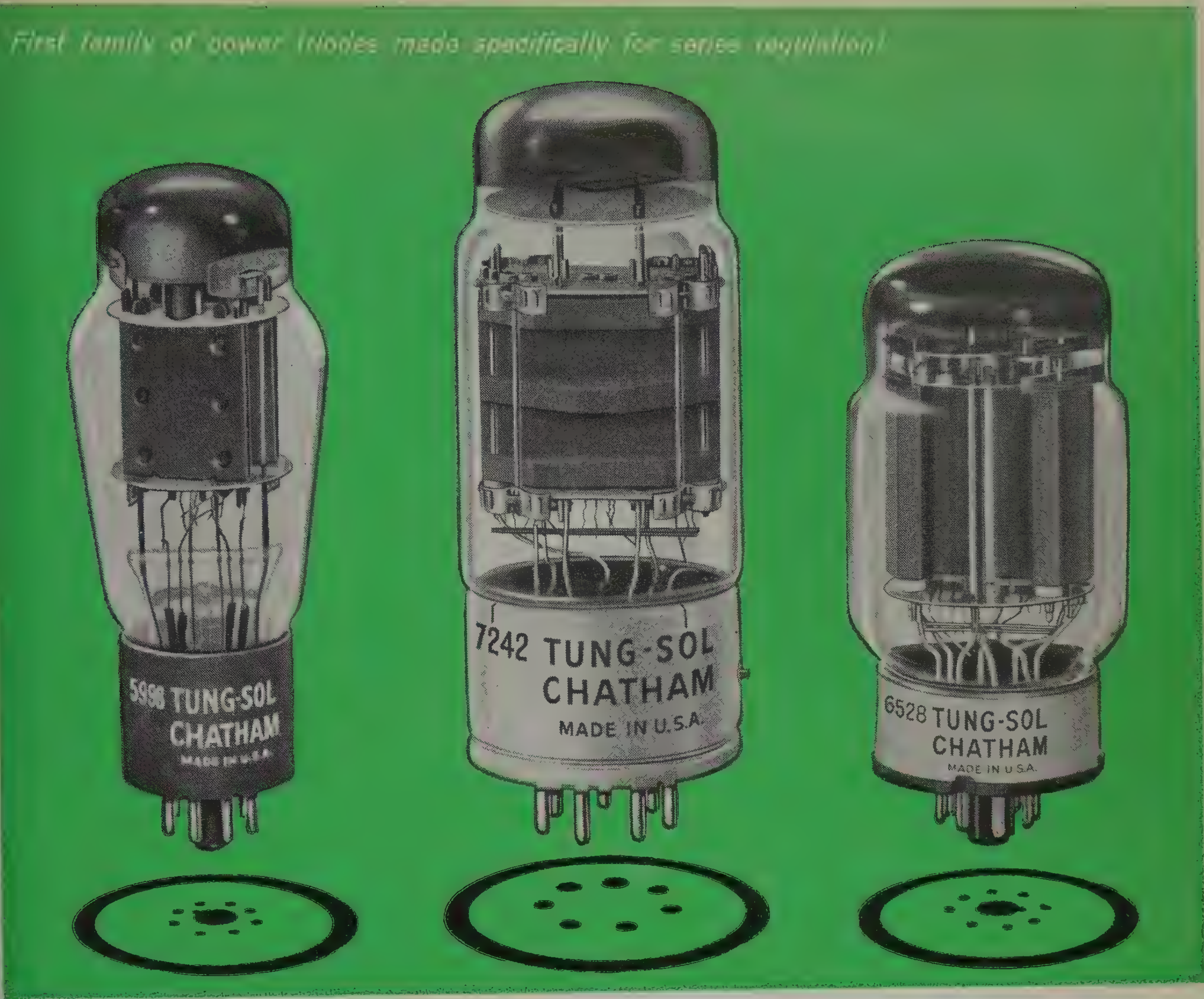
"Competitive Fixed-Price Creativity—Product Development," J. W. Leas, RCA.

Instrumentation

Long Island—December 16

"The Application of Computers to Railroad Freight Classification Yards," T. Dosch, Reeves Instrument Corp.

(*Continued on page 104A*)



Tung-Sol/Chatham power triode family covers every series regulator need!

Now designers can specify a premium quality Tung-Sol/Chatham tube for all series regulator sockets. Tung-Sol/Chatham's family of power triodes — the first designed and produced specially for series regulator service — meets all design requirements and assures maximum reliability and life at all times.

Types include the new 100 Watters, 7241 and 7242, medium mu or low mu-high current. 12 or 26 Volt heater versions available on most types. All embody sturdy construction features that contribute to overall ruggedness and long hours of heavy-duty operation.

Compare the ratings below against your particular application! If you desire complete data sheets . . . or you have a specific design problem, contact us today! We'll be glad to give whatever assistance we can. Just write: Tung-Sol Electric Inc., Newark 4, N. J., Commercial Engineering Offices: Bloomfield and Livingston, N. J., Culver City, Calif., Melrose Park, Ill.

TUNG-SOL®

TYPICAL VALUES				
	Total Plate Current	Range of Tube Voltage Drop	Minimum Tube Drop	Grid Voltage Swing
5998	200 ma	80 v	45 v	20 v
6528	400	65	70	10
7242	600	80	70	13

PERTINENT CHARACTERISTICS PER TUBE				
	Max. Plate Current	Max. Plate Voltage	MU	Gm
5998	280	275	5.5	28,000 umhos
6528	600	400	9.0	74,000 umhos
7242	900	400	9.0	111,000 umhos

TUBE TYPES BY PLATE DISSIPATION RATINGS			
Total Plate Dissipation	26 to 30 W	60 W	100 W
Low Mu	6AS7G, 6082 6080WA, 7105	6336A 6394A	7241
Medium Mu	5998	6528	7242

AN IMPORTANT NEW PRODUCT ANNOUNCEMENT FROM HUGHES!

PARAMETRIC AMPLIFIER DIODES

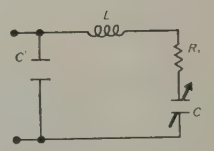
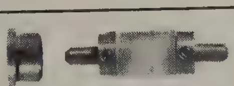
FOR LOW-NOISE MICROWAVE AMPLIFIERS

Now Hughes Products brings you high performance parametric amplifier diodes at a price in the same range as good microwave mixer crystals. These Hughes diodes have been designed to solve your problems associated with low-noise parametric amplifiers, modulators, frequency converters, harmonic generators, electronic tuners, switches, etc., at microwave as well as at lower frequencies.

Used in a 3000 Mc high gain parametric amplifier with both signal and idler channels as inputs, these diodes have produced at room temperature in the laboratory a noise temperature of 100°K above absolute zero. Noise temperatures of 50°K above absolute zero were obtained when diode was cooled by liquid nitrogen.

The Hughes Parametric Amplifier Diodes are available in two rugged, hermetically sealed versions. One has a miniaturized glass package (type HPA 2800); the other has been adapted to a conventional microwave package (type HPA 2810). Both are hermetically sealed in glass and have the same cutoff frequency.

TECHNICAL SPECIFICATIONS AND DATA:

Package (actual size)	C′	C @ zero bias (nominal)	cutoff frequency* (nominal)	L (nominal)	V_S** Min.	Nom.	Equivalent Circuit
HPA 2800 	0.1$\mu\mu$f	2.5$\mu\mu$f	70KMC	4mμh @ 1KMC	5V	7V	
HPA 2810 	0.2$\mu\mu$f						

*At breakdown voltage **Breakdown voltage (10μA point)

CAPACITANCE vs. BIAS VOLTAGE

Reverse Bias Voltage	Capacitance C
0 V	2.5 $\mu\mu$f
3 V	0.76 $\mu\mu$f
7 V	0.60 $\mu\mu$f

Address inquiries to:
Hughes Products, Semiconductor Marketing Dept.,
P. O. Box 278, Newport Beach, California.

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SEMICONDUCTOR DEVICES • STORAGE AND MICROWAVE TUBES • CRYSTAL FILTERS • OSCILLOSCOPES • RELAYS • SWITCHES • INDUSTRIAL CONTROL SYSTEMS

Professional Group Meetings

(Continued from page 100A)

Washington—January 5

"The U. S. Lunar Probes," C. E. Brown, Nat'l Aeronautics and Space Admin.

"The U. S. Lunar Probes," J. F. Clark, NASA; R. Brady, Chief, Communications Branch, Signal Corps Res. and Dev., also spoke on the Project SCORE communications results.

Medical Electronics

Boston—November 25

"The Coding of Sensory Information by Insect Nerves," K. Roeder, Tufts College.

San Francisco—December 2

"Airborne Medicine," N. P. Thompson, Palo Alto Medical Research Foundation.

San Francisco—January 6

"Hospitals in the Electronic Age," Moderator, Mark S. Blumberg, Stanford Res. Institute; Panel, Marke Berke, Director, Mt. Zion Hospital, and San Francisco; R. V. A. Lee, M.D., Director, Palo Alto Medical Clinic; A. J. Morris, vice-president, Levinthal Electronic Prods., Palo Alto.

Microwave Theory and Techniques

Baltimore—November 19

"Microwave Propagation—Earth-Bound and Space," D. E. Sukhia, The Martin Co.

Baltimore—January 28

"Broadside Mattress Array Using Surface Wave Transmission Modes," H. W. Cooper, Westinghouse Air Arm Div.

Los Alamos-Albuquerque—January 14

"Microwave Refractometry," D. C. Thorn, Univ. of N. M.

Omaha-Lincoln—January 16

"Solid State Masers," R. H. Hoskins, Hughes Res. Lab.

Omaha-Lincoln—January 21

"The KD Distance Relay in the Microwave Scheme," J. E. Hagburg, Westinghouse Elec. Corp.

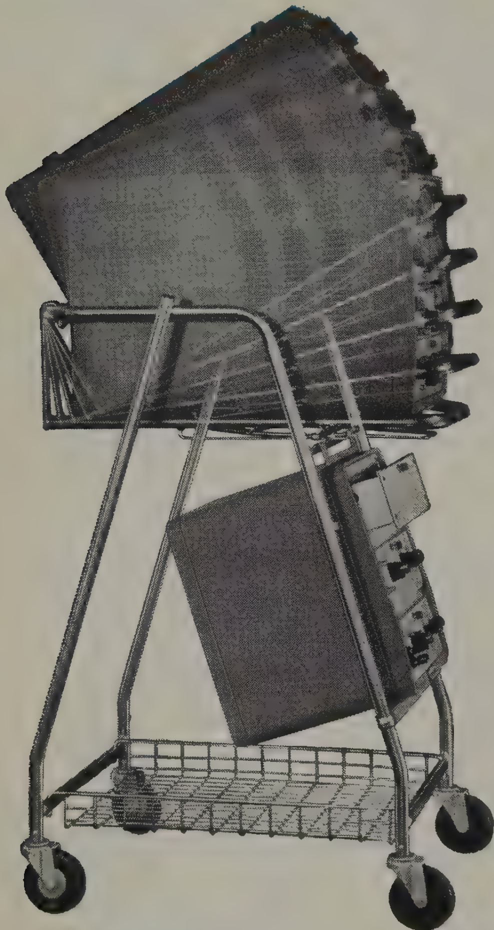
Military Electronics

Central Florida—November 25

"Radiation Belt and Other Radiation Problems," S. F. Singer, Univ. of Maryland.

(Continued on page 108A)

No stoop, no squint, no painful nagging backache*



Buy this Testmobile and tilt your 'scope so you can read it!

Obsoleting all previous concepts in one brilliant breakthrough, *-hp-* engineers have achieved the *ultimate device*—the revolutionary 115A Oscilloscope Testmobile. Employing the radical Supermarket Cart principle (first described 1906 by A. and P.) *-hp-* 115A *actually tilts an oscilloscope so you can read it, and lets you push it from place to place!* Scope may be tilted up to 30° in 7½° increments; heavy chromed tube steel construction; big, locking, rubber-tired wheels; removable bottom basket; size 40″ high x 23″ wide x 29″ deep, folds for shipment or storage; lightweight, only 28 lbs., $80.

Still further probing the Unknown, *-hp-* engineers achieved the *-hp-* 116A Storage Unit and 117A Storage Drawers. The 116A is a sophisticated cube known as a "box." It holds up to 3 plug-in units for *-hp-* 150A/AR 'scopes; prevents dust and elbows in the circuitry. Yours for $22.50. The 116A also holds up to three 117A drawers which in turn hold tools, solder, components and bubble gum. *-hp-* 117A, a modest $10.

*with thanks to our friends at Philco and Anacin

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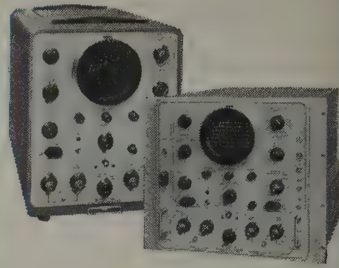
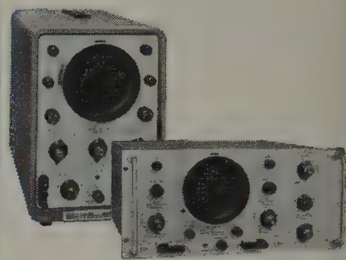
Cable "HEWPACK" • DAvenport 5-4451
Field engineers in all principal areas

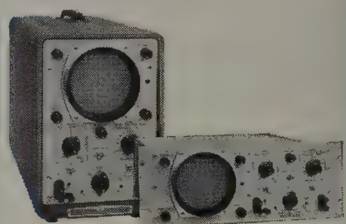
Besides Testmobiles, *-hp-* makes oscilloscopes, too!



***-hp-* 150A/AR - to 10 MC**
Automatic trigger, direct-reading; plug-ins providing dual trace or differential input; or high amplification. *-hp-* 150AR (rack) $1,200. *-hp-* 150A (cabinets) $1,100.



***-hp-* 130B/BR - to 300 KC**
1 mv sensitivity, similar X/Y amplifiers, direct reading, automatic trigger, X5 magnifier, balanced on 6 most sensitive ranges. *-hp-* 130B (cabinet) or 130BR (rack), $650.



***-hp-* 120A/AR - to 200 KC**
Sweeps 1 μsec/cm to 0.5 sec/cm; X5 sweep magnifier, automatic trigger, high sensitivity calibrated vertical amplifiers, regulated power supplies. *-hp-* 120AR (rack mount, 7″ high) or 120A (cabinet) $435.

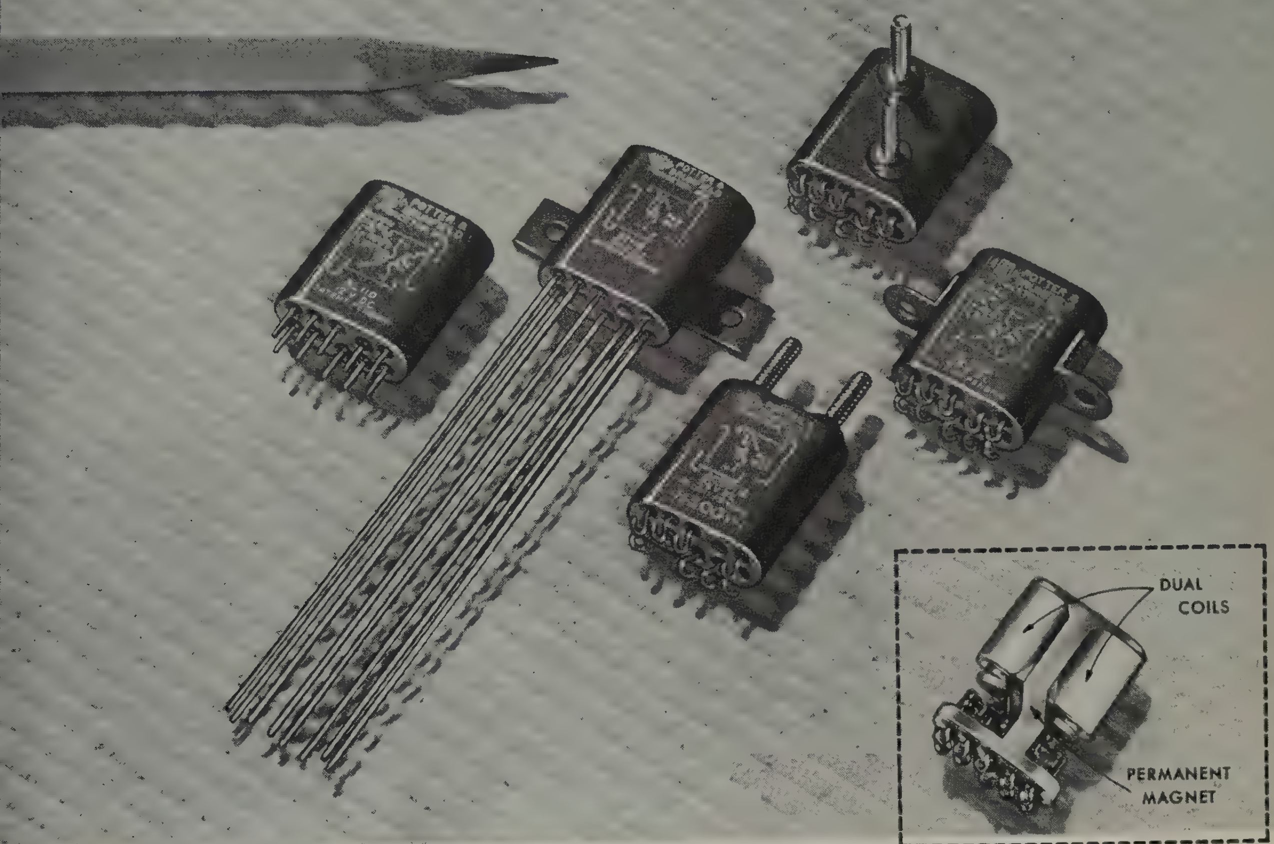
Data subject to change without notice. Prices f.o.b. factory

hp over 300 instruments for measuring speed and accuracy

P&B MICRO-MINIATURE RELAYS LEAD IN

performance

SHOCK: 100g* VIBRATION: 30g to 2000 cps*

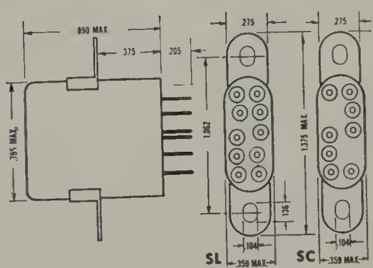


*NO CONTACT OPENING

New P&B crystal-case size relays, the SC and the SL (magnetic latching), show amazing shock and vibration capabilities. They absorb shocks of 100g and vibrations 30g to 2000 cps. without contact openings!

A highly efficient magnetic structure utilizing a permanent magnet makes possible at least twice the contact pressure found in DPDT relays of comparable size. One watt of power for three milliseconds operates either relay. Transfer time is unusually fast —0.5 milliseconds maximum.

For more information, contact your P&B sales engineer, or write Potter & Brumfield, Princeton, Indiana.



SL—dual coil latching relay. Operates on a 1 watt, 3 ms. pulse at nominal voltage. Permanent magnet latch locks the armature in either position.

SC—non-latching relay with series-connected dual coils. Operates on approximately 1 watt at nominal voltage. Coils must remain energized to hold the armature in the operate position.

SC and SL Series Engineering Data

GENERAL:
Insulation Resistance: 10,000 megohms, min.
Breakdown Voltage: 1,000 V. RMS.
Shock: 100g.
Vibration: 30g 55 to 2000 cps.; 0.195″ max. excursions from 10-55 cps.
Temperature Range: —65° C. to + 125° C.
Weight: 15 grams without mounting bracket.
Operate Time: 3 MS. max. with 550 ohm coil @ 24 V. DC. (SL: 630 ohm coil at 24 V. DC).
Transfer Time: 0.5 MS max.
Terminals: (1) Plug-in for microminiature receptacle of printed circuit board.
(2) Hook end solder for 2 #24 AWG wires.
(3) 3″ flexible leads.
Enclosure: Hermetically sealed.

CONTACTS:
Arrangement: 2 Form C.
Material: Optional
Load: 2 amps. @ 28 V. DC, resistive; 1 amp @ 115 V. 60 cycles AC, resistive.
Pressure: SC—16 grams min.; SL—20 grams min.

COIL:
Power: Approx. 1.0 watt at Nominal Voltage.
Resistance: SL—40 to 10,000 ohms; SC—35 to 20,000 ohms.
Duty: Continuous.

MOUNTINGS:
Bracket, stud and plug-in.

P&B STANDARD RELAYS ARE AVAILABLE AT YOUR LOCAL ELECTRONIC PARTS DISTRIBUTOR

POTTER & BRUMFIELD INC.

SUBSIDIARY OF AMERICAN MACHINE & FOUNDRY COMPANY

Professional Group Meetings

(Continued from page 104A)

Los Angeles—November 18

"Lunar Probe Instrumentation," A. W. Newberry, Space Electronics Corp.

"Lunar Probe Telemeter Design," Y. Shibuya, Space Technology Labs.

San Diego—October 28

"Application of Pulsed Doppler to Airborne Radar Systems," W. W. Maguire, Hughes Aircraft Co.

Nuclear Science

Dayton—October 2

"Air Force Nuclear Engineering Test Facility," K. P. Howard, U. S. Air Force.

Dayton—November 13

"The Air Force Institute of Technology Subcritical Reactor," W. L. Lehman, U. S. Air Force Inst. of Tech.

Production Techniques

Boston—January 20

"Engineering changes and their effect on production and sales," panel discussion: R. W. Waters, T. Spotswood, and E. Hampson.

(Continued on page 110A)

Professional Group Meetings

(*Continued from page 108A*)

San Francisco—January 27

"Testing Operations on Limited Run Production Electronic Equipment," E. B. Edberg, Varian Assoc., and W. J. Rolly, Ampex Corp.

Reliability and Quality Control

Florida West Coast—November 12

"Effect of Normal and Non-Normal Distribution on Systems Reliability," C. Wyatt, General Electric Co.

Florida West Coast—January 21

"High Reliability at Low Cost," B. L. Weller, Vitramon, Inc.

Los Angeles—January 19

"Application of Statistics to Reliability Engineering," H. G. Romig, Hoffman Electronics Labs.

Space Electronics and Telemetry

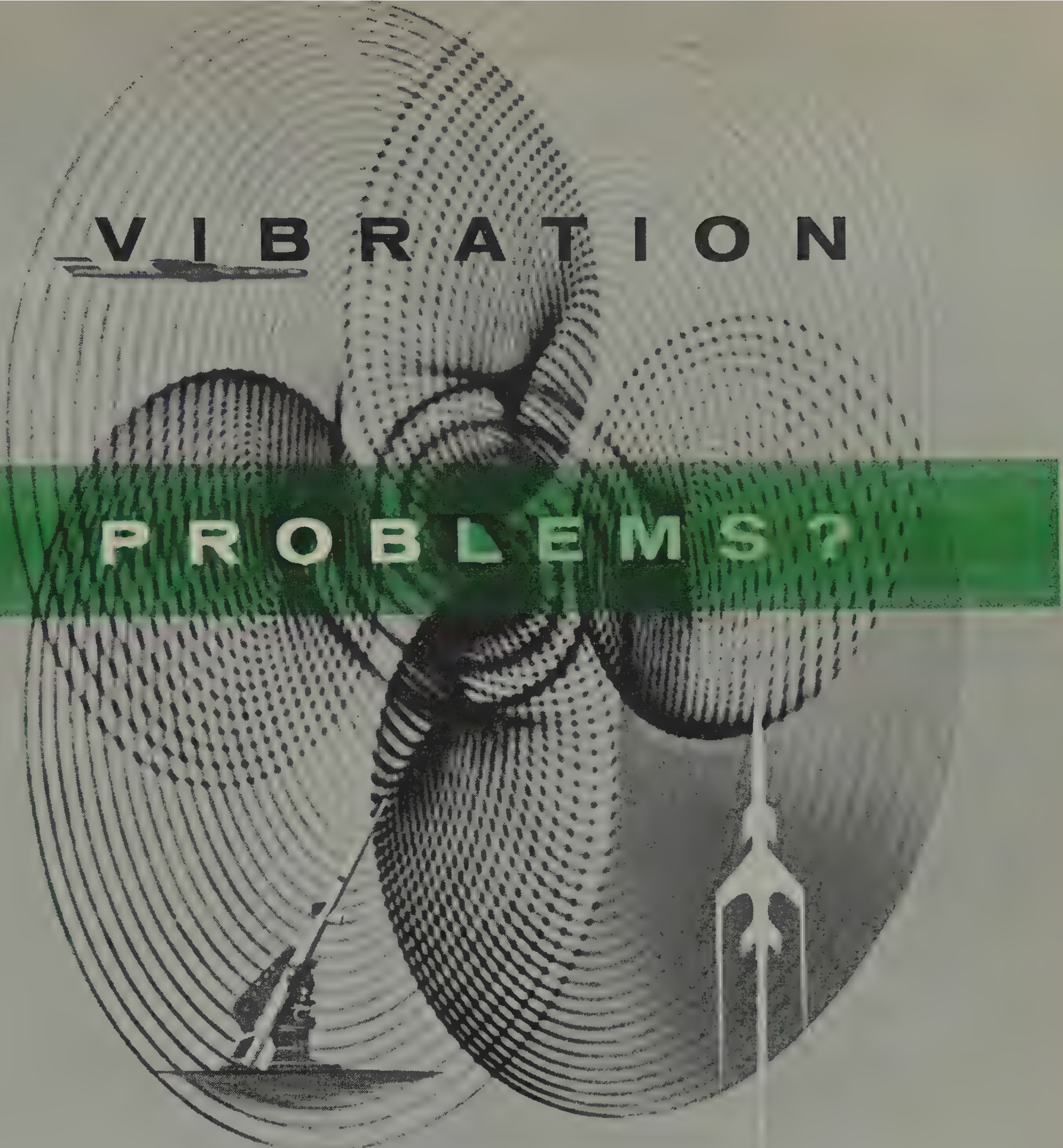
Albuquerque-Los Alamos—September 16

Business meeting—opening of ballots to elect new officers.

(*Continued on page 112A*)

VIBRATION

PROBLEMS?



CALL OUT CANNON PLUGS

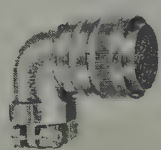
Improvements are constantly being made in Cannon connectors to give you maximum reliability in circuitry under constantly increasing highly critical vibration conditions. New lines... improvements on the MS-E design... include the EX, the CT, the EA, and the EB Series.

You'll find Cannon vibration and moisture-proof resilient-insert lines extensive and complete, in hundreds of carefully designed layouts. Interfacial sealing. Improved grommets and grommet followers. Styles with extra strong coupling nuts. Telescoping rubber bushings. Strong clamps. Means for safety wiring. Grounding lugs. Every facility to give solutions to vibration, moisture condensation, flashover, and corona problems.

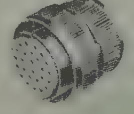
Vibration-proof Series include MS-E, CA-F, EA, EB, EX, and CT. Write TODAY for full information.

MS-E (Series CT)
Socket Assembly
Ask for
MS-E Catalog

CA 3100 F
Wall Mounting Receptacle
Ask for
AN/MS Catalog



MS-E (Series ME)
63 Inserts Available
Ask for
MS-E Catalog



EX 06
Plug Assembly
Ask for
EX Catalog

CANNON PLUGS
WHERE RELIABILITY IS THE 5th DIMENSION
AWARENESS OF PROBLEM · INTEGRITY OF INTENT · DESIGN EFFORT · MANUFACTURING CONTROL

Write for 44-page, 2-color Plug Guide, Bulletin CPG-3

Please Refer to Dept. 377

CANNON ELECTRIC CO., 3208 Humboldt St., Los Angeles 31, Calif. Factories in Los Angeles, Salem, Mass., Toronto, Melbourne, London. Manufacturing licensees in Paris and Tokyo. Representatives and distributors in all principal cities.

Professional Group Meetings

(*Continued from page 110A*)

Philadelphia—January 21

"Problems of Attaining Reliability in Missiles and Space Vehicles," E. D. Karmiol, Gen. Elec. Co.

Note: The originally planned speaker could not be in Philadelphia because of travel difficulties.

VEHICULAR COMMUNICATIONS

Florida West Coast—November 19

"Microwave System Planning and Applications," L. Elmore, Motorola Communications and Electronics, Inc.

Positions Wanted

By Armed Forces Veterans

In order to give a reasonably equal opportunity to all applicants and to avoid overcrowding of the corresponding column, the following rules have been adopted:

The IRE publishes free of charge notices of positions wanted by IRE members who are now in the Service or have received an honorable discharge. Such notices should not have more than five lines. They may be inserted only after a lapse of one month or more following a previous insertion and the maximum number of insertions is three per year. The IRE necessarily reserves the right to decline any announcement without assignment of reason.

Address replies to box number indicated, c/o IRE, 1 East 79th St., New York 21, N.Y.

MANAGER—ENGINEER

Manager of electronic or communication engineers—Mature responsible executive; Substantial electronic-communication background; fully developed qualities of leadership, resourcefulness and judgment. Able to inspire team work and high morale; outstanding achievements in organization, coordination and development of personnel. Box 1085 W.

APPLICATION AND LIAISON ENGINEER

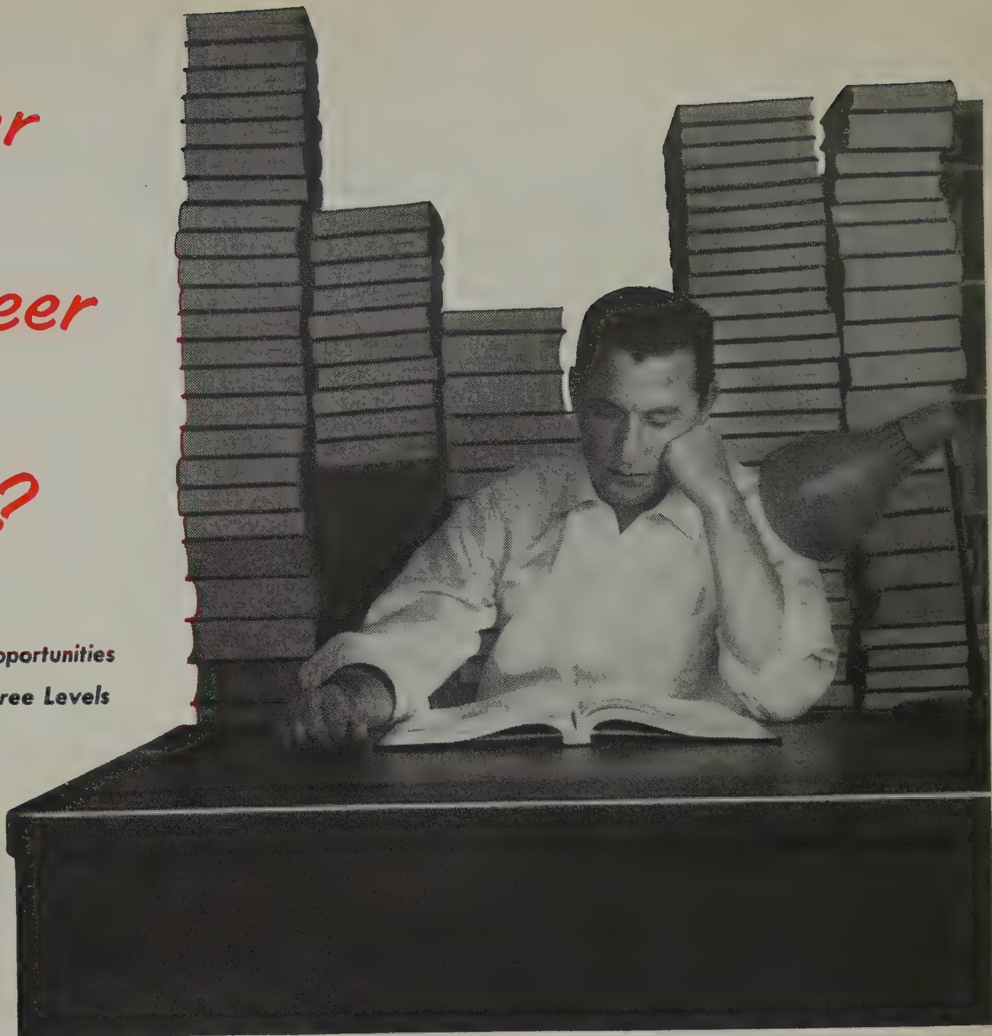
Reserve Signal Officer formerly with combat developments office, U.S. Army Special Warfare School, desires position as application and liaison engineer with company interested in the development of special communications and electronic equipment for military application. Box 1086 W.

EDUCATOR—AUTHOR—ENGINEER

MS. School administrator, proven record in course development electronics technology at Community College and institute level, desires administrative or training position in school or in electronics or allied industry. Age 47, married, 1 child. Box 1087 W.

(*Continued on page 114A*)

How far can an engineer go at AC?



Special Advanced Education Opportunities
Develop Career Potential on Three Levels

If you are a graduate engineer in the electronics, electrical or mechanical fields, or if you have an advanced degree in mathematics or physics, you can work on AC's famous AChiever inertial guidance system or a wide variety of other electro-mechanical, optical and infra-red devices. You can also prepare for promotion and enhance your professional status through AC's free comprehensive training programs—the finest "in house" programs available.

Program A—for recent graduate engineers—gives you a solid foundation in the theory and application of inertial guidance systems and servomechanisms. You attend classes three hours per day for four months, all on company time.

Program B—for experienced engineers—consists of upgrading studies in inertial guidance, servomechanisms, environmental problems, engineering math and physics, plus advanced state-of-the-art courses. Time? —during working hours or evenings.

Program C—for all engineering supervisors—involves management training developed by a team of AC executives and University of Chicago industrial relations experts. Sixty one-half-hour sessions give you a solid grounding in management techniques.

These thoroughly practical courses—taught by university professors or recognized AC specialists—constitute AC educational "extras." AC offers them in addition to their educational assistance programs for men who wish to study for advanced degrees in nearby universities.

AC invites you to compare these advanced training offerings with those of any other company. Get complete details today. Just write the Director of Scientific and Professional Employment: Mr. Robert Allen, Oak Creek Plant, Dept. E, Box 746, South Milwaukee, Wisconsin.

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AC

AC SPARK PLUG
THE ELECTRONICS DIVISION
OF GENERAL MOTORS

Positions Wanted

By Armed Forces Veterans

(Continued from page 112A)

ELECTRONIC ENGINEER

Graduated RPI, BEE 1955. 2 years project and field engineering of lt. wt. radar. 1/Lt. USAF with 2 years experience in ECM and large scale digital computer programming. Desires project engineering work in the eastern U.S., preferably New York area. Age 24, married. Box 1088 W.

MARKETING ENGINEER

BSEE. and MBA. 6 years design and supervisory experience. Seeks marketing position with a small electronics firm which shows growth potential. Box 1097 W.

SALES ENGINEER

Experienced in both jobber and OEM phases. Currently performing as exclusive representative to large OEM account involved in both commercial and government programs. Desires to relocate in Southern California. Box 1098 W.

ELECTRONIC ENGINEER

10 years of varied professional experience including circuit design, component manufacture and systems analysis in the fields of communications, missile guidance and weapons system evaluation. Desires responsible and challenging position in small or medium size San Francisco Bay area. MSEE. plus additional graduate work in administration engineering. Senior Member IRE; Ex-Signal Corps Officer; Top secret clearance. Box 1099 W.

(Continued on page 118A)

Months before Pearl Harbor, JPL had tested America's first liquid rocket engines using spontaneously igniting propellants. By April 1942, a simple nitric acid-aniline propulsion system was designed into and successfully tested in an A-20-A Bomber for a jet-assisted takeoff. For high-altitude atmosphere research purposes, JPL then used the hypergolic liquid rocket system in the WAC CORPORAL. Placed as a second stage on a V-2 rocket, this became the BUMPER WAC rocket that established a World's altitude record of 242 miles in February 1949.

At the request of U.S. Army Ordnance, the Jet Propulsion Laboratory now began to develop a long-range guided ballistic missile system, incorporating the proven, smooth-burning light-weight acid-aniline system. These achievements sparked the development of a whole series of rocket vehicles. In 1954, the Army accepted the JPL developed CORPORAL, which became America's first tactical guided ballistic missile system; its accuracy exceeded design requirements.

Under the direction of the National Aeronautics and Space Administration, the experienced Jet Propulsion Laboratory research and development team is now working on storable, high-performance hypergolic liquid propulsion systems with which space vehicles may soon orbit the moon and planets.

jpl

CALIFORNIA INSTITUTE OF TECHNOLOGY

JET PROPULSION LABORATORY

A Research Facility of the National Aeronautics and Space Administration

PASADENA, CALIFORNIA

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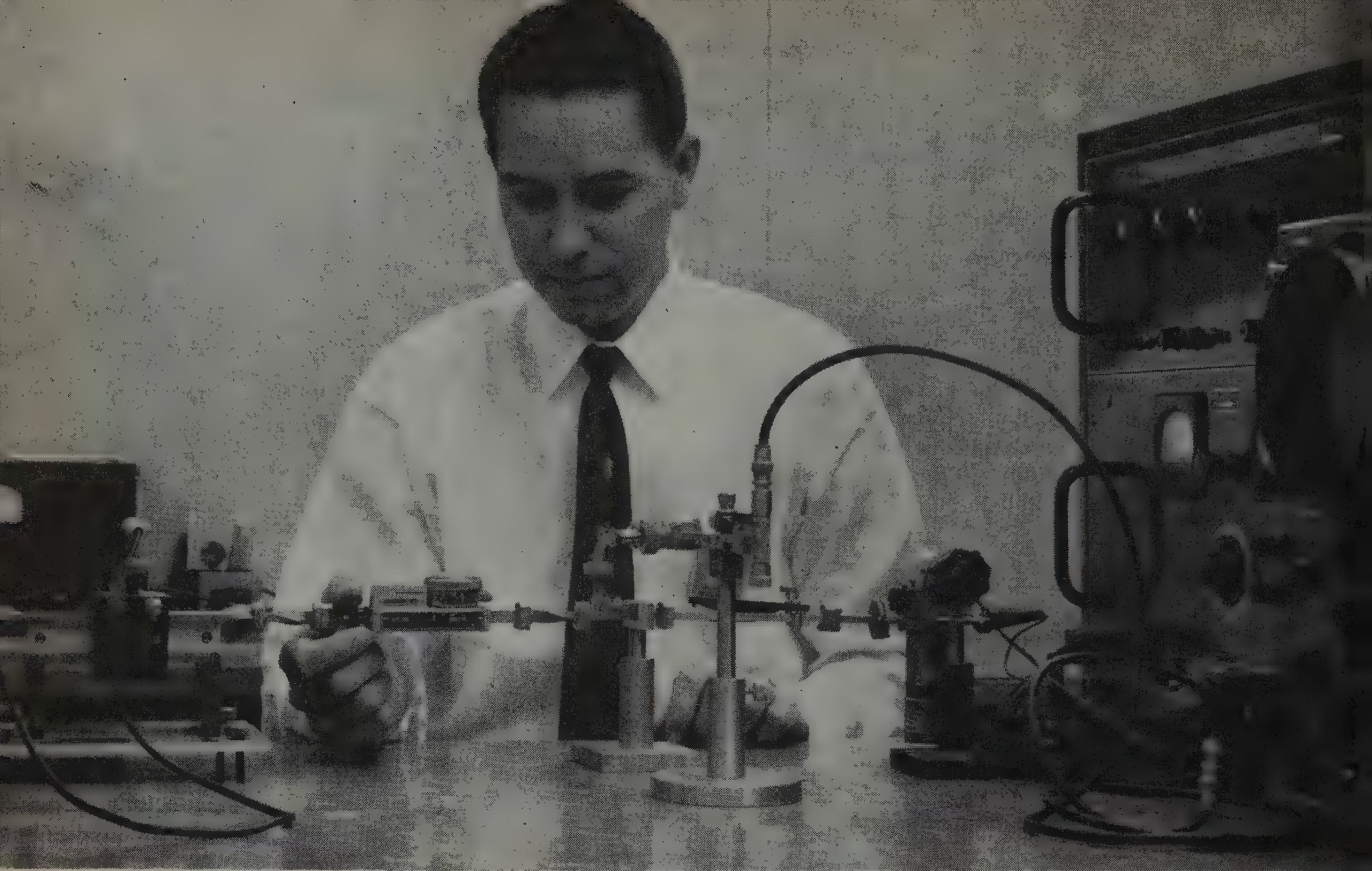
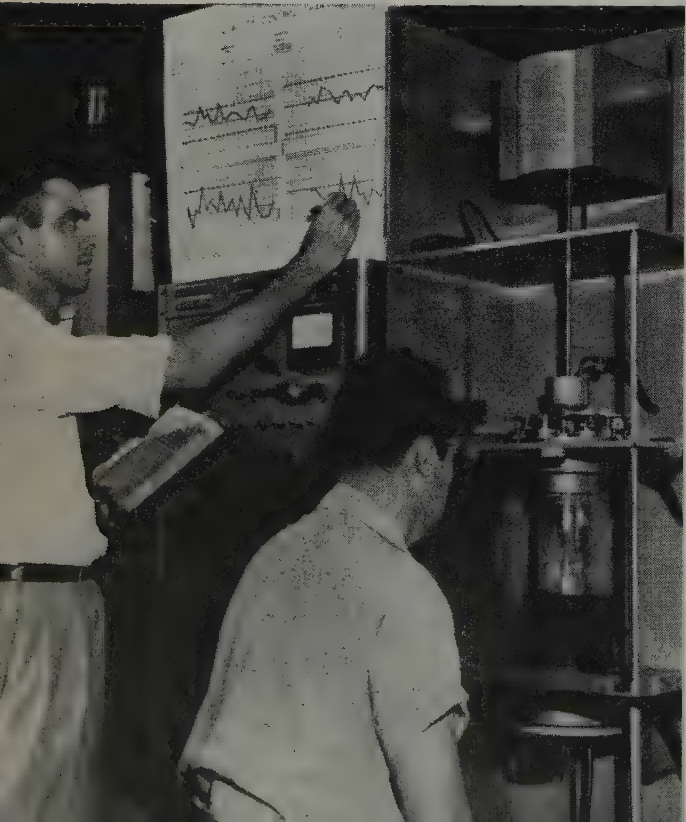
ELECTRONIC, MECHANICAL, CHEMICAL, PROPULSION, INSTRUMENTATION, MICROWAVE, AERONAUTICAL AND STRUCTURAL ENGINEERS






seisMAC

Positions Wanted

By Armed Forces Veterans

(Continued from page 114A)

DEVELOPMENT ENGINEER

15 years experience, particularly strong in electro-mechanical field and analog computers. Also has done hydraulic, optical, and electronic development. Career interest in management and technical direction of R&D activities. Box 2000 W.

TECHNICAL CONSULTING

Ph.D. MIT 1953. Assignments in microwave heating, other electronic applications to biological and food problems, technical surveys, diversification studies, and information research. Registered professional engineer, inventor, teacher and author. Box 2011 W.

COMPUTER SCIENTIST

Computer Scientist with 17 years general electronics experience. Large experience in translation of Russian electronics literature. Will accept single-shot or steady outside translation work up to 75,000 words per month. Box 2012 W.

ENGINEER

Engineer with BS. degree wishes to relocate to small town or suburb on west coast. 10 years design and development experience including television, Airborne VHF communication and RF and AF transistor circuit design. Top 10% of graduation class. Now located with A-1 manufacturer in midwest. Box 2013 W.

(Continued on page 120A)

ATLAS: MIGHTY MISSILE WITH A MIGHTY MISSION

In this young missile age, the mighty Atlas already commands a unique and dominant role. A proven weapon system, it is also our only large, tested booster and the only known vehicle to steer itself into orbit. Atlas will lift the first manned capsule into space and many exploration programs of the future include this superb missile in their planning.

Newest program to be released at Convair-Astronautics is **PROJECT CENTAUR**—design, construction and testing of a high energy, upper stage rocket. Boosted by Atlas, this vehicle will be able to place in orbit a satellite weighing several thousand pounds.

ELECTRONIC ENGINEERS at Convair-Astronautics participate in space projects which are consistently at the most advanced state of the art. Positions are available in many areas and at various levels of experience. There are also immediate openings for engineers and scientists in electrical, mechanical and structural design, systems design and analysis, propulsion, test laboratories, field test, engineering writing, computer programming, research engineering, and other specialties. Our engineering representatives will be conducting **INTERVIEWS** in these and other cities soon:

ALBUQUERQUE, BALTIMORE, BOSTON, BUFFALO, CHICAGO, CLEVELAND, DAYTON, DENVER, DETROIT, EL PASO, FT. WAYNE, INDIANAPOLIS, KANSAS CITY, LOS ANGELES, MINNEAPOLIS, NEW ORLEANS, NEW YORK, PHILADELPHIA, PITTSBURGH, ST. LOUIS, SPRINGFIELD, MASSACHUSETTS, AND WASHINGTON, D.C.

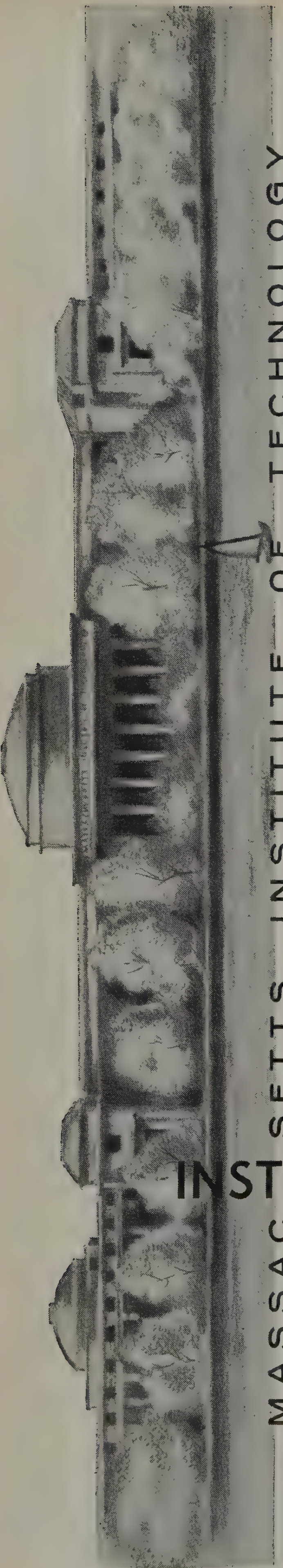
We invite qualified engineers and scientists to send a detailed resume at once so advanced arrangements can be made for a confidential interview. Write to Mr. G. N. McMillan, Engineering Personnel Administrator, Department 130-90.

CONVAIR/ASTRONAUTICS

CONVAIR DIVISION OF

GENERAL DYNAMICS

5510 KEARNY VILLA ROAD, SAN DIEGO, CALIFORNIA



Research and Development for the Seeking Mind

ELECTRONICS ENGINEERS

ALL LEVELS

The Laboratory, with its staff of 850 employees, is primarily engaged in the conception and perfection of completely automatic control systems necessary for the flight and guidance of aircraft, missiles and space vehicles.

R and D opportunities exist in:

System Design & Theoretical Analysis
Astronautics
High Performance Servomechanisms
Power Supplies
Magnetic Amplifiers
Analog and Digital Computers
Electro-mechanical Components
Transistor Circuitry
Printed Circuitry
Environmental Instrumentation & Evaluation
Research, Design & Evaluation of Gyroscopic Instruments
Computer Programming
Simulator Studies
Classical Mechanics
Optical Instrumentation
Pulse Circuitry
and in many other areas.

INSTRUMENTATION LABORATORY

WRITE:

I. R. SAMUELS
Director of Personnel

M.I.T. INSTRUMENTATION LABORATORY

68 Albany Street
Cambridge 39, Massachusetts

*Graduate courses may be taken for credit while earning full pay.

U. S. Citizenship Required

Positions Wanted

By Armed Forces Veterans

(Continued from page 118A)

ENGINEERING MANAGER OR ADMINISTRATOR

BSEE., MSEE., MBA., PE. 12 years of microwave and television design, systems and administration experience. Presently Engineering Director. Seeking high level responsible position. Location secondary to opportunity. Box 2014 W.

RADIO ENGINEER

BSEE. University of Calif., Berkeley; Graduate work University of Calif. and Johns Hopkins, MS. anticipated 1959. Lt. USNR; age 28, married, 1 child. Experience varied industrial work as student, 3 years naval operational, 2 years electronics instructor Naval Academy. Seeks position in communications, circuit design or general electronics leading to position as systems engineer. Prefers southern New England or Washington, D.C. area. Box 2015 W.

ADVERTISING AND PUBLIC RELATIONS

12 years intensive experience in program planning, budgeting and administration; product, market and media analysis; technical advertising, sales promotion and publicity copy; company-agency liaison, Government, industry, community and press relations. Licensed radio operator/mechanic with marketing background in electronic, nucleonic, aviation and allied in-

(Continued on page 125A)

ELECTRONICS ENGINEERS

(Senior)

- Develop test equipment, methods and procedures for determining conformance of complex electronic equipment with Company and Air Force specifications.
- Contact various electronic manufacturers to determine requirements for purchase of equipment as required to test and simulate flight operations of radar, navigation, fire control & similar electronic systems.
- Also develop procedures for use of electronic test equipment. Conduct investigatory work and recommend corrective action and changes. EE degree required. Quality control experience desired.

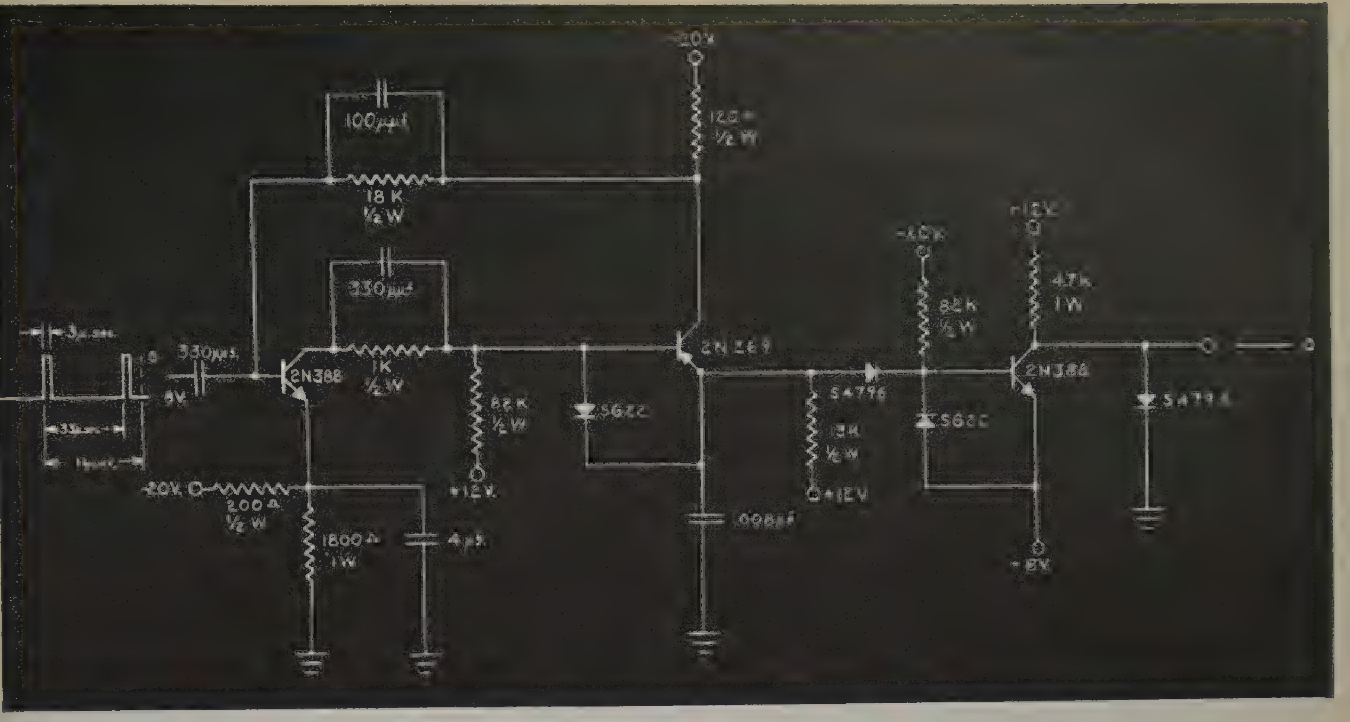
Salary commensurate with ability.

In addition to other advantages, Republic offers a comprehensive benefit program among the finest in industry.

Send resume in confidence to

Mr. W. Walsh
Employment Office

REPUBLIC AVIATION
Farmingdale, Long Island, N.Y.



FOR COMPUTER AND DATA-PROCESSING ENGINEERS...

***NCR** has select openings at the New National Engineering-Research Center in Dayton, Ohio*

HERE ARE THE TYPES OF ENGINEERS WE NEED:

COMPUTER ENGINEERS:

****Senior Systems Engineers*—Strong Theoretical and Design Knowledge in Electronic Engineering, including familiarity with electro-mechanical digital machines. Prefer experience with commercial application of digital-processing equipment, will consider scientific or defense application. Operational experience a distinct asset. Advanced degree desired.

Your Work at NCR—analyze and direct product improvement of digital computers.

****Senior Circuit Designers*—experienced in the design, development and analysis of transistorized computer circuits, including application of magnetic cores to high-speed memories.

Your Work at NCR—opportunities involving decision making concerning reliability, cost and component selection are offered.

****Senior Circuit and Logical Designers*—similar experience and duties as noted for Senior Circuit Designers plus evaluation and de-bugging arithmetic and control areas of computer systems.

DATA-PROCESSING ENGINEERS:

****Senior Electronic Design Engineers*—experienced in the development of logical design using standard computer elements.

Your Work at NCR—to evaluate and design transistorized circuits including design transistorized regulated power supplies and circuitry related to decimal to binary coding.

WHERE YOU WILL WORK...

at NCR's NEW Engineering Research Center, Dayton, Ohio.

You'll be working under the most stimulating and advanced R and D facilities with broad creative freedom in the engineering field which is *yours.*

HOW DO I APPLY?

Simply send your résumé to: Mr. K. D. Ross, Professional Personnel Section H, The National Cash Register Company, Dayton 9, Ohio.

* TRADEMARK REG. U.S. PAT. OFF.

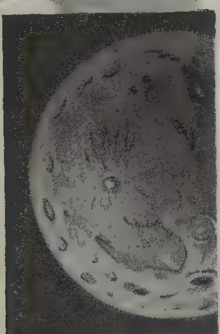
National*

VERSATILE DATA PROCESSING
ADDING MACHINES · CASH REGISTERS
ACCOUNTING MACHINES · NCR PAPER

THE NATIONAL CASH REGISTER COMPANY, DAYTON 9, OHIO

ONE OF THE WORLD'S MOST SUCCESSFUL CORPORATIONS

75 YEARS OF HELPING BUSINESS SAVE MONEY



advanced airborne flight data system for **MANNED FLIGHT INTO SPACE** again emphasizes $\boxed{\text{creative}}$ fulfillment for **SPERRY ENGINEERS**

When the hypersonic X-15 rocket aircraft attempts a manned flight into space, it will be equipped with one of the most advanced airborne flight data systems ever devised — a development of Sperry Gyroscope Company.

This highly advanced Sperry inertial system is one of a long series of creative accomplishments by Sperry engineers.

Back of these achievements: a unique combination of working conditions . . . *on* important, diversified assignments . . . *with* America's foremost engineers . . . *for* a company famed as an "engineer's firm". A company which has grown steadily for almost a half century, and whose engineers have grown with it, as proved by over 2,600 Sperry employees who are 15-year men. When you go with Sperry, you grow with Sperry. *If you're interested, not in an engineering job but in an engineering CAREER, check Sperry.*

Openings in many specializations including:

INERTIAL NAVIGATION SYSTEMS

DOPPLER NAVIGATION	RADAR RECEIVERS
RADAR TRANSMITTERS	INFRARED SYSTEMS
PULSE CIRCUITS	SERVO SYSTEMS
ELECTRONIC PACKAGING	GYROS & ACCELEROMETERS
TRANSISTOR CIRCUITS	MICROWAVE ANTENNA DESIGN

Confidential Interviews
Contact Mr. J. W. Dwyer, Employment Manager
Saturday Interviews 8 A.M.-1 P.M.
Arranged by Appointment

SPERRY
GYROSCOPE COMPANY
Division of Sperry Rand Corp.
Great Neck, Long Island, N. Y.
Fieldstone 7-3665

Positions Wanted

By Armed Forces Veterans

(Continued from page 120A)

dustries. BS.—Engineering Management, MBA.—Industrial Marketing, Ph.D. candidate—Advertising. Age 39. Desires company or agency affiliation with broader responsibilities, leading to marketing management. Box 2016 W.

ELECTRONIC ASSISTANT

Married veteran; Associate of Applied Science, Degree in Electrical Technology. 1st class Radio Telephone license, Signal Corps Communications school training. Seeks interesting and challenging position in sales or as a sales correspondent, as technician, as engineer assistant or as broadcast technician in communications or electronics industries. Northeast area preferred. Box 2017 W.

MANAGEMENT—ENGINEER

Broad background in operational requirements for military systems, particularly in airborne electronics. Also in organization and management. Desires challenging position of responsibility, preferably in mid-west. BS. (USNA), MS. in engineering electronics. Naval aviator. Age 38. Available July 1959. Box 2018 W.

Positions Open

The following positions of interest to IRE members have been reported as open. Apply in writing, addressing reply to company mentioned or to Box No. . . . .

The Institute reserves the right to refuse any announcement without giving a reason for the refusal.

Proceedings of the IRE
1 East 79th St., New York 21, N.Y.

ELECTRONICS ENGINEER

For R&D on an entirely new class of components employing thin metal films. Some experience required in microwave components or deposition of thin metal or inorganic films. Write Vice Pres., Engineering, Box 1088.

FACULTY POSITIONS

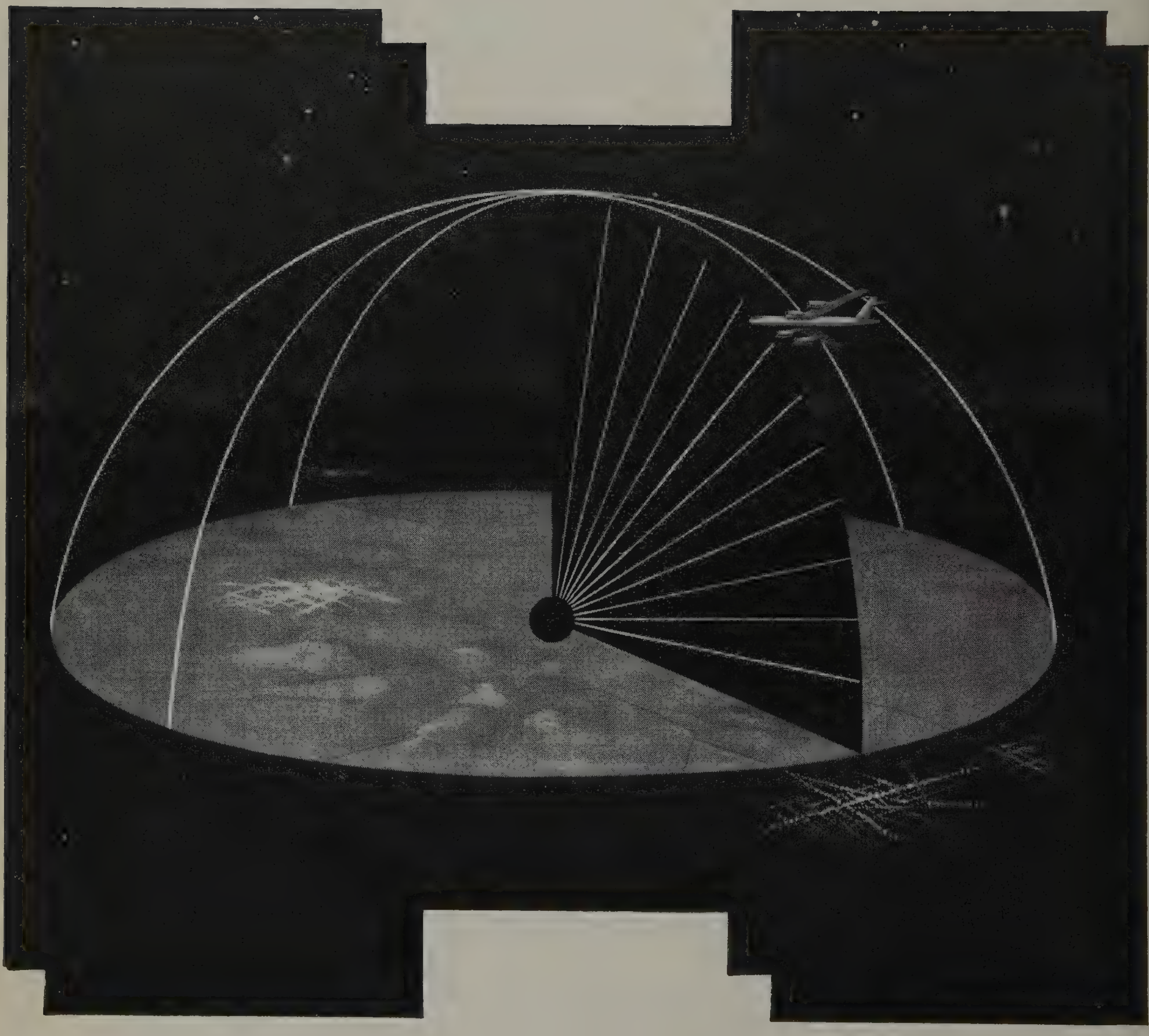
Assistant Professor positions open for persons with Ph.D. degree or equivalent in solid-state electronics, computers, networks, controls, microwaves and others. Part-time teaching associate positions also available for persons with MS. degree or equivalent industrial experience who wish to work for Ph.D. degree. Research and teaching assistantships for other graduate students. Write to Chairman, Dept. of E.E., University of California, Berkeley 4, Calif.

SENIOR ELECTRONIC SYSTEMS ENGINEERS

These are project level positions for direction of teams developing major electronic instrumentation systems, and automatic data processing systems. Complete responsibility for all phases. Require broad electronic system and circuit design experience. Attractive salaries and bonus. Send resume to W. H. Hammer, Director of Research and Development, Land-Air, Inc., P.O. Box 394, Holloman, N.M.

(Continued on page 128A)

THE UMBRELLA



THAT NEVER LEAKS

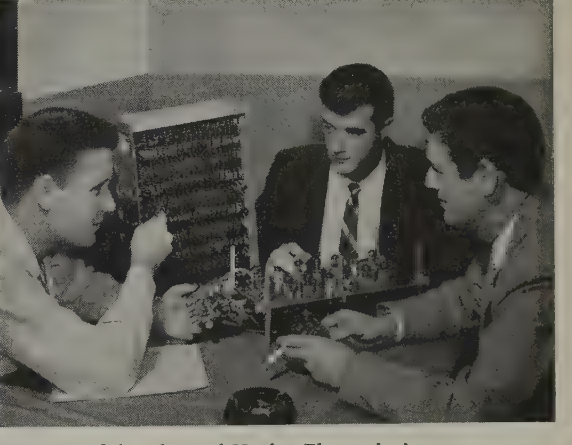
To achieve umbrella-like radar protection Hughes engineers at Fullerton, California, have developed systems which position radar beams in space by electronic rather than mechanical means. These unique three-dimensional radar systems are digitally programmed to instantaneously detect high-speed enemy aircraft, even at low altitude.

Other defense systems under development at Hughes in Fullerton are Data Processors which monitor the movement of hundreds of aircraft, store the information and assign defense weapons; radars with beams capable of detecting and tracking missiles; and new radar systems for installation on surface and subsurface naval vessels.

Other Hughes activities are delving into similarly advanced areas of electronics. Engineers at Hughes Research & Development Laboratories are probing into the effects of nuclear radiation on electronic equipment, studying advanced microwave theory and applications, and examining communication on a spatial scale. Applying this advanced type of creative engineering to commercial projects is the task of engineers at the Hughes Products activity.

The highly advanced and diversified nature of Hughes projects offers creative engineers and physicists the opportunity to build a rewarding career in a progressive and expanding environment.

Research & Development Engineers use REAC computing equipment as an aid in such complex problems as systems simulation.



Reliability of the advanced Hughes Electronic Armament systems can be insured only with the equally advanced test equipment designed by Hughes El Segundo engineers.

Newly instituted programs at Hughes have created immediate openings for engineers experienced in the following areas:

Semiconductors	Development Engineering
Field Engineering	Systems Analysis
Microwave Engineering	Industrial Systems
Industrial Dynamics	Circuit Design
Communications	Components Engineering
Digital Computers	Electron Tubes

Write in confidence, to Mr. Tom Stewart, Hughes General Offices, Bldg. 6-E4, Culver City, California.

Creating a new world with ELECTRONICS

HUGHES

HUGHES AIRCRAFT COMPANY
Culver City, El Segundo,
Fullerton and Los Angeles, California
Tucson, Arizona

Positions Open

(Continued from page 125A)

SENIOR COMMUNICATIONS ENGINEER

Experienced in microwave, radio, and wire line communications. Capable of design, cost estimating, preparation of economic studies, writing and interpreting specifications for equipment units or entire systems, to supervise communications studies and coordinate work with other departments and companies. Write, giving resume of education, experience and qualifications, including picture and telephone number. Box 1090.

RESEARCH AND DEVELOPMENT

Openings for electrical engineers interested in the design and development of electronic and electro-mechanical devices for industrial clients. Degree in E.E. or physics required. Industrial experience helpful. Please send resume to Mr. John C. Stowe, Arthur D. Little, Inc., 35 Acorn Park, Cambridge 40, Mass.

INDUSTRIAL CONSULTING

Opening for an electrical engineer interested in market research, primarily for the electronics industry. Experience in industry and/or M.B.A. required. Please send resume to Mr. John C. Stowe, Arthur D. Little, Inc. 35 Acorn Park, Cambridge 40, Mass.

ASSISTANT OR ASSOCIATE PROFESSOR

Assistant or Associate Professor with M.Sc. or Ph.D. degree. To teach communications or electronics courses and direct advanced degree

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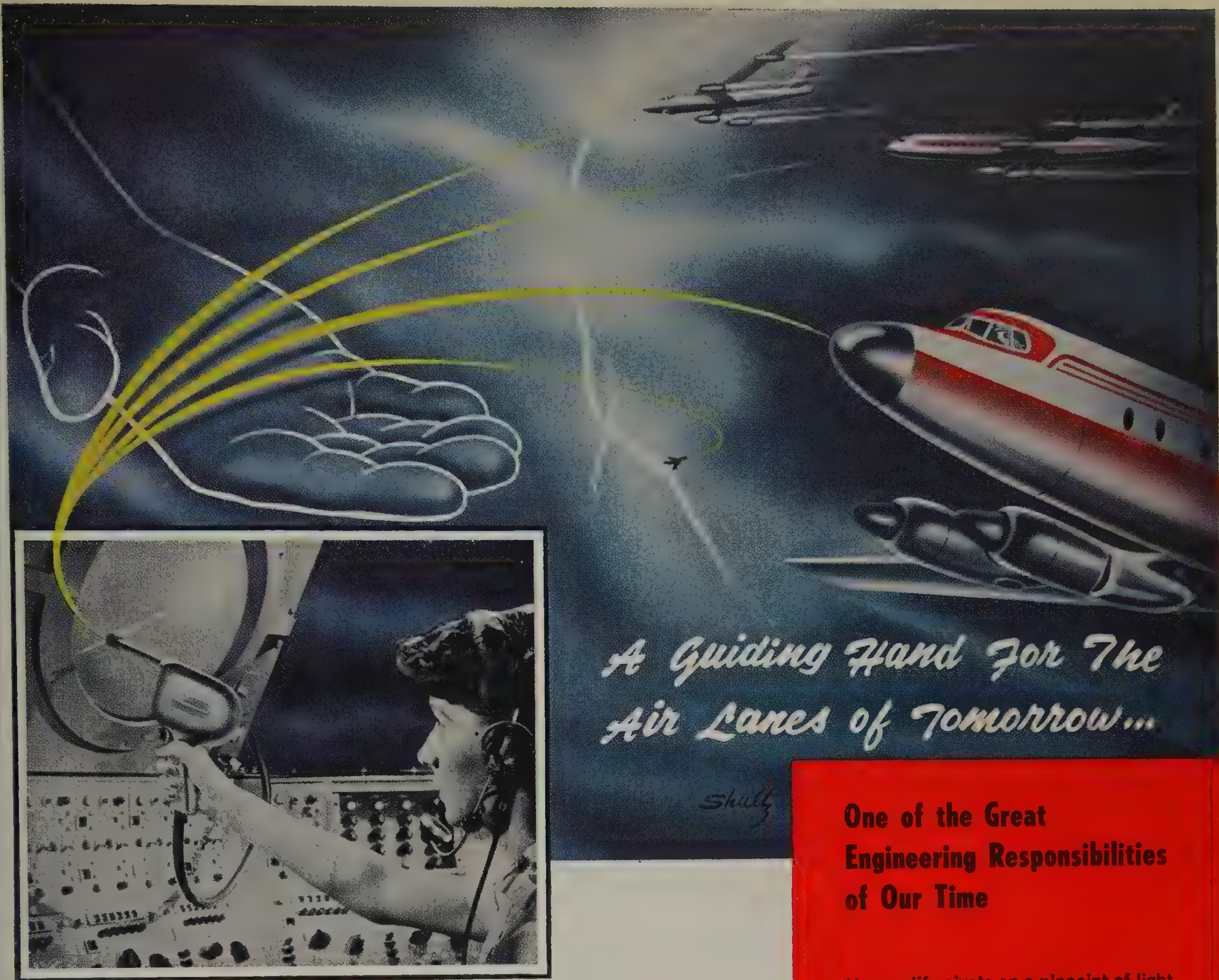


One of the Great Engineering Responsibilities of Our Time

Human life pivots on a pinpoint of light in a desolate sea of blackness . . . held in place by a thread of precise mathematical orientation.

Up to the present, there have been too many variables in the equation of air safety; and the dawn of the jet era has suddenly made present air traffic control systems obsolete.

Avco/Crosley is absorbed in the vital task of eliminating these variables from the air safety equation . . . with the AN/GSN-11 Air Traffic Control System, a Program now under development at our Cincinnati facility, in cooperation with the Air Force.

The guiding hand of this Air Traffic Control System will soon be expediting military air traffic with the utmost precision . . . Course, altitude, rate of descent will all be accounted for in the final answer.

Join the men who will cross these new frontiers . . . you'll be glad that you had a share in building a brighter, safer future for American aviation.

OPPORTUNITIES FOR ENGINEERS....

In Air Traffic Control

A chance to own a share in America's future security.

Air Traffic Control — Experience with systems comprising radar search, computers, display and vehicle command. System synthesis, analysis and equipment development experience needed.

Telemetry and Guidance—Design experience on components and sub-systems in advanced military equipment for drone and weapon control.

Radar Surveillance — System and equipment design experience in high power heavy ground radar installations necessary. Acquaintance with associated data handling, computers and decision equipment required.

Technical Writing — Maintenance manuals, instruction books, progress reports, technical proposals and presentations.

Correspondence regarding these positions should be addressed to:

Director, Scientific and Technical Personnel

CROSLEY DIVISION—AVCO

1329-G Arlington Street, Cincinnati 25, Ohio

Positions Open

(Continued from page 128A)

candidates in communications or electronics beginning Sept. 1, 1959. Opportunities for research. Salary depends upon qualifications. Write Chairman, Dept. of E.E., University of Nebraska, Lincoln 8, Neb.

MAGNETRON CATHODE ENGINEER

Engineer or scientist with minimum of 2 years experience in magnetron cathodes. Work in non-military research and development. Stable employment with an aggressive company. Send resume to Franklin Mfg. Co., 65-22nd Ave. N.E., Minneapolis 18, Minn. Att: Sam Kellough.

CRITICAL VACANCIES AT GRIFFISS AIR FORCE BASE, ROME, N.Y.

Electronic Scientist	GS-7	\$4980 pa
Electronic Engineer	GS-5	\$4490 pa
Industrial Engineer	GS-5	\$4490 pa

INSTRUCTORS

For those interested in teaching fundamental electrical and electronic subjects. Openings for fall semester beginning September 1959. Salary depends on qualifications. New building and equipment. Opportunity for advancement and further study. 9 month academic year. Part-time evening session work available. Health insurance and pension plan. Send complete resume to Professor I. L. Koscow, Dept. Head, Electrical Technology, Staten Island Community College, 50 Bay St., Staten Island 1, N.Y.

TECHNICAL WRITERS

Teleregister designs and develops some of the world's largest data processing systems. We have immediate openings for both Junior and Senior technical writers to prepare a variety of brochures and manuals. Moving expenses paid, many company benefits. If you would like to join a growing company, send your resume to Paul N. Callender, The Teleregister Corp., 445 Fairfield Ave., Stamford, Conn.

TEACHING—RESEARCH

Full year appointments available. Research needs presently in field of electronic countermeasures with other fields possible. Teaching in new undergraduate program with strong engineering science emphasis and E.E. graduate (MS.) program. MS. or Ph.D. required. 9 month salary range presently \$5000-\$9000. Arrangements can be made for combined teaching-research or for full research and teaching alternate semesters or years. Apply A. T. Murphy, Head, Dept. of E.E., University of Wichita, Wichita, Kan.

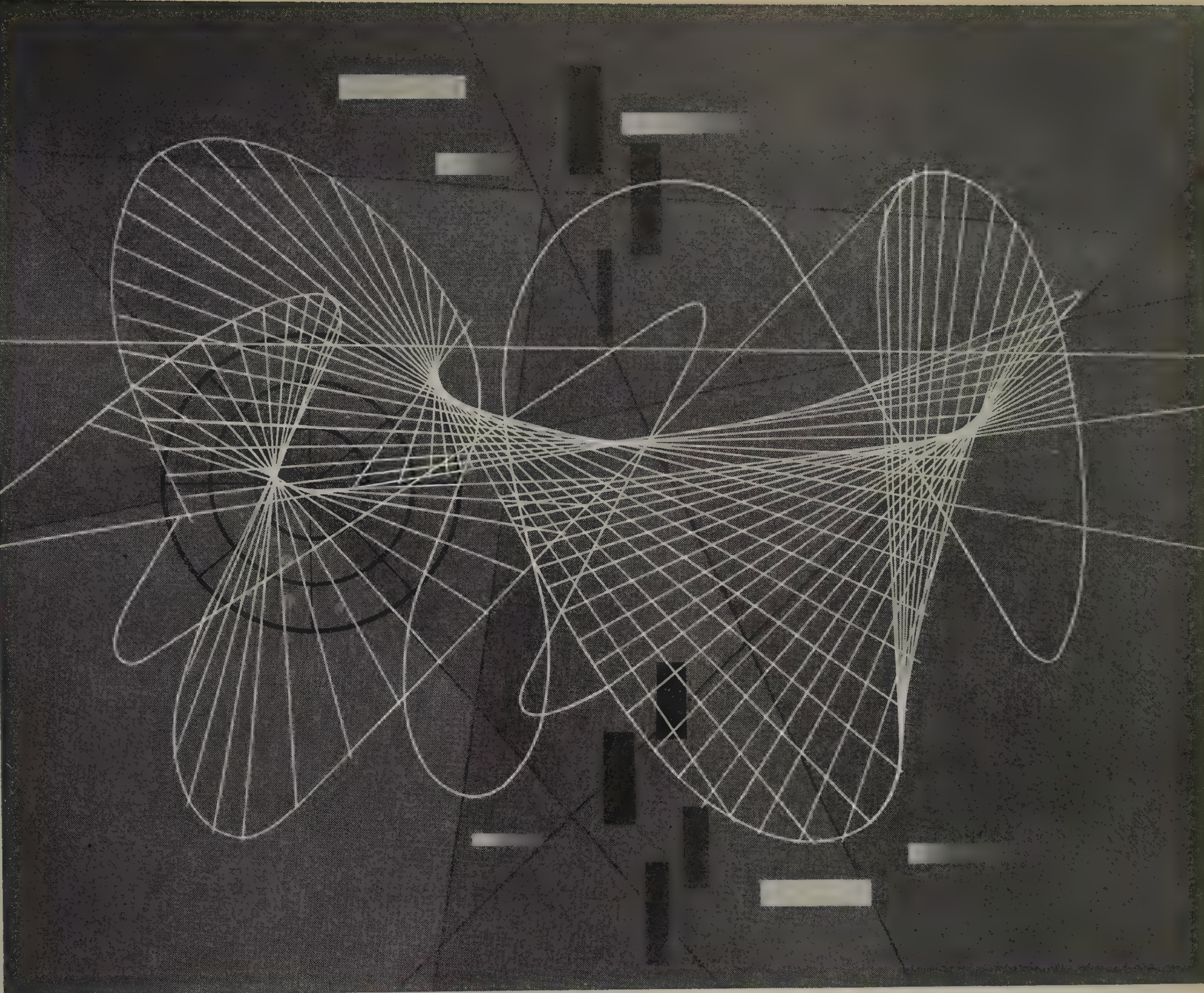
TEACHING—ELECTRICAL ENGINEERING

An unusual opportunity exists for a recent graduate with MS. or Ph.D. Stimulating teaching assignments in a modern curriculum offer a unique combination of desirable professional experience. Salary range for 9 months is \$5000 to \$7000. Attractive full year contract available. Location is in the midwest at a medium size university. Send complete resume to Box 1092.

COMPUTER DEVELOPMENT ENGINEERS

E.E.'s or physicists in the areas of: memory storage systems; transistor and magnetic core logical systems; logical design; analysis and

(Continued on page 132A)



Positions Open

(Continued from page 130A)

programming; and displays. Salaries: $12,000-$16,000. Write Mr. Aston, Witty-Polon Management Consultants, 25 East 73 St., New York 21, N.Y.

TECHNICAL—RESEARCH

Technical gentleman with knowledge of research market to join group of scientists as stockholder and/or associate. Duties: Contact government and industry research facilities regarding research proposals. Manhattan Physical Research Group, Inc., 556 West 191 St., New York 40, N.Y.

ELECTRICAL ENGINEERING PROFESSORS

Fast growing engineering school needs 3 electronic-electrical engineers able to teach circuits and electronics, with a specialty in communications, industrial electronics or computers. MS. degree plus industrial experience is the minimum requirement; a Ph.D. is desirable. Salary and rank depends on experience. The department has just moved into a new building and needs aggressive men who wish to develop curricula and laboratories. Apply: A. D. Graves, Dean, Los Angeles State College, Los Angeles 32, Calif.

TECHNICAL WRITERS

Burroughs Corp. Military Field Service Div. is expanding long term Atlas & Sage programs in data processing, creating growth positions for technical writers with experience writing to military specifications on electronic equipment.

(Continued on page 135A)

Positions Open

(Continued from page 132A)

Must have a good background in electronic theory. Qualified men will work in our modern facility, located in an attractive Phila. suburb. Send resume to Mr. A. J. Bellace, Employment Supervisor, Dept. R-789, Burroughs Corp., Military Field Service Div., Burroughs Dr., Radnor, Pa.

PROFESSORS

Applications are invited for one position as Assistant Professor and one as Associate Professor in any field of Electrical Engineering. Positions combine half-time teaching and half-time research. Applicants must have Ph.D. or demonstrate equivalent qualifications by published research. Send full details to Chairman, Div. of Engineering, Brown University, Providence 12, R.I.

SALES MANAGER

Electronic Instruments, Commercial Products Dept. of long established progressive engineering and manufacturing organization located on Long Island, N.Y. Must have engineering education with substantial successful sales management experience in electronic test and control instruments. Dept. operates through manufacturing representatives selling to prime contractors, government agencies, industrial organizations, and educational institutions. State experience, education and salary expected. Box 1093.

TEACHING POSITIONS

The Electrical Engineering Dept. of the City College of New York has several positions available on the teaching staff beginning Sept. 1959. Rank and salary commensurate with qualifications and experience. Opportunity for graduate study. Address inquiry to Prof. H. Taub, Dept. of Electrical Engineering, The City College, Convent Ave. and 139 St., New York 31, N. Y.

NEWS New Products

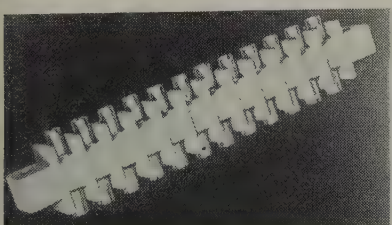
These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 40A)

losses of less than 1 db per harmonic have been observed in many experiments. More information is available from the firm, request Catalog 59V.

Terminal Blocks

A line of heavy-duty terminal blocks designed by the Bureau of Ships and covered by applicable MIL specifications, is announced by **Kulka Electric Corp.**, 633–643 S. Fulton Ave., Mount Vernon, N. Y.



Representative of the line is Type 7TB12 made to Navy drawing 9000

(Continued on page 136A)

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 135A)

S6506B, 73214, Rev. H. It provides for feedthru connections at top and bottom, and comes in several different lengths and number of terminals. It is moulded of glass-filled Alkyd plastic (Type MAI-60) as per MIL-M-14E. The moulded-in threaded studs are of manganese-bronze. It is supplied with slotted brass nuts made to specifications, packaged separately or supplied assembled, as desired.

Details regarding these Navy-type terminal blocks and other types, are available on request from the firm.

Frequency Converter

Power Sources, Inc., of Burlington, Mass., leading manufacturers of transistorized power sources, announces the availability of the new Model PS6001 Frequency Converter.

The unit, which meets all requirements of military specifications for ground equipment, has an input of 120 volts, 60 cps, and an output of 115 volts, 400 cps, 250 watts square-wave.

The converter will operate at any temperature between −30°C and +52°C, and is not damaged by temperature extremes to −65°C and +85°C. It meets all shock and vibration specifications for military ground equipment.

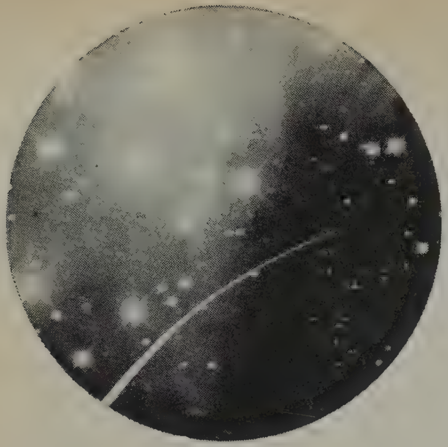
The case measures $10\frac{1}{2} \times 5 \times 7$ inches, and weighs 20 pounds. A standard military type AN connector is supplied.

The PS6001 is currently in production for use by the Air Force and is available in quantity for other military or commercial applications. Further details may be obtained by writing to the firm.

Miniature Beam Switching Tubes

New miniature Beam Switching Tubes in both shielded (Type BD 316) and unshielded (Type BD203) versions have been announced by **Burroughs Corp., Electronic Tube Div.**, P.O. Box 1226, Plainfield, N. J., for low voltage applications requiring minimum size and weight. The new tubes operate at transistor voltages, and

(Continued on page 140A)

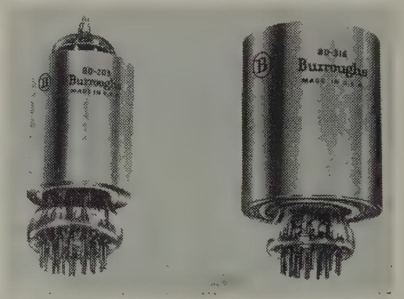


NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 136A)

retain all of the characteristics of regular Beam Switching Tubes with the added advantages of reduced size and weight. Physical dimensions of the tubes have been reduced by 35 per cent without affecting inherent reliability. Miniature shielded tubes may be used in direct contact with each other without magnetic interference thus providing high-packing density for airborne applications.



Miniature Beam Switching Tubes are 10 position high vacuum electronic tubes used for electronic distribution, counting, coding and decoding, sampling and conversion. Each tube may replace twenty or

(Continued on page 143A)

ıese manufacturers have invited PROCEEDINGS ɛaders to write for literature and further technical ıformation. Please mention your IRE affiliation.

(Continued from page 140A)

ıore transistors or other components in lectronic circuitry as the ten output posiions, each with automatic memory and igh impedance switching, are controlled y a single electron beam. Tubes can be rranged to switch in sequence or at ranlom, can be preset to any position, interonnected as distributors to any number f positions, and used for a wide variety f electronic generating, switching and lriving operations.

For complete information write to the irm.

Transformer Catalog

PCA Electronics, Inc., 16799 Schoenɔorn St., Sepulveda, Calif., has published ı new 24-page catalogue titled *"Pulse Transformers"* and designed to assist engiıeers in the application of transformers to their specific needs.

Complete with many tables, charts, and schematics this manual covers a brief history of low-level pulse transformers, their measurements, specifications, applications, interchangeability, dielectric ratings, manufacturing, and other data.

Also included is information on some of PCA's 2,000 standard design transformers (available from stock), as well as case types and specifications data to aid the engineer who requires transformers of custom design and manufacture.

Pulse Transformers is available on request by writing to the firm.

Pfeffer Heads Struthers-Dunn

Announcement of the appointment of John L. Pfeffer as president, effective the first of the year, has been made by **Struthers-Dunn, Inc.,** relay manufacturers of Pitman, N. J. Formerly assistant general manager, Pfeffer has been associated with the company since leaving the Navy in 1946. He succeeds the retiring president, Henry W. Pfeffer, who headed the firm for the past 20 years.

John L. Pfeffer is also vice-president of the National Association of Relay Manufacturers.

IPC Extends Memberships

Users of printed circuits are now eligible for membership in the **Institute of Printed Circuits.** The announcement came from W. J. McGinley, President of Methode Manufacturing Corp., and President of the IPC, following a recent meeting of the Institute in Chicago.

The IPC, which has recently published a book entitled "How to Design and Specify Printed Circuits" is a national trade

(Continued on page 144A)

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 143A)

association. The objective of this association is to accelerate the growth of the industry through the development of standards; through activities which will advance the technology; and through promoting the attributes of printed circuits.

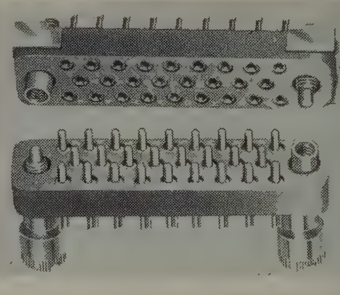
The decision to widen the scope of membership in the IPC was made to accommodate several large manufacturing companies who use printed circuits in their equipment and who expressed a desire to be a part of IPC programs. It was the consensus of both the users of printed circuits and the present membership of the IPC that the technological advancement of the printed circuit industry could be greatly accelerated by working together on activities of common interest.

Allied Membership in the IPC is available to companies and corporations in the electronics and electrical field that use printed circuits in their products. McGinley reports that dues for Allied Membership are $500.00 per year.

Further information can be obtained by contacting the Institute of Printed Circuits 27 E. Monroe St., Chicago 3, Ill.

Subminiature Connectors

A new series of subminiature precision connectors was made available recently by **U. S. Components, Inc.,** 454 E. 148th St., New York 55, N. Y.



A stainless steel reinforcing retainer is provided under each screwlocking element to remove all torque stresses from the molded bodies, avoiding breakage. Positive re-entrancy of the male pins is assured each time by a specially flanged guide female contact. Self-alignment action is also assisted by provision of wider countersink on upper end of contact. The connectors are available in total contacts of 7, 11, 14, 20, 26, 32 or custom configurations may be obtained to meet specific requirements.

The new SMI-C Subminiature Series precision connectors are ideal for critical environmental conditions and extremes of military applications. For further details, write to the firm.

(Continued on page 146A)

News from Raytheon's Semiconductor Division...

ELECTROLYTIC SLICING— This engineer is slicing a germanium crystal by electrolytic means. Up to now semiconductor wafers have been formed by mechanical processes, such as cutting with diamond saws or lapping with abrasive powders. The resulting mechanical damage to the critical surfaces reduces the quality and effectiveness of finished semiconductor devices. Electrolytic slicing of crystals, producing surfaces which are free from the mechanical damage resulting from other methods, is one of the many pioneering activities initiated and carried forward by the scientists and engineers of Raytheon's Semiconductor Division.

THE PLACE FOR THE MAN WHO IS GROWING FASTER THAN HIS ASSOCIATES

There are openings at Raytheon's Semiconductor Division for scientists and engineers with semiconductor experience and a desire to find more room for personal and career growth. Opportunities exist in the following areas:

Device Design and Development
Material Development
Mechanization
Circuit Design
Application Engineering

You are invited to explore the advantages for yourself in associating with Raytheon's Semiconductor Division. Write to Mr. Allen I. Moorhead, RAYTHEON MANUFACTURING COMPANY, Semiconductor Division, 150 California Street, Newton 58, Massachusetts.

The place for the man who is growing faster...
SEMICONDUCTOR DIVISION of RAYTHEON

Excellence in Electronics

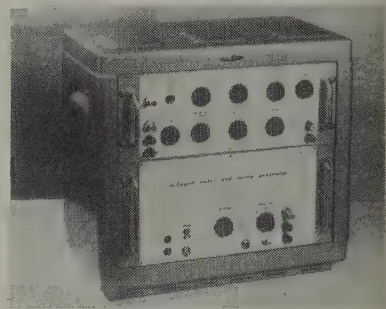
NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 144A)

Pulse Generator

Marconi Instruments, 111 Cedar Lane, Englewood, N. J., announces a new pulse generator, Cintel Model 3351, with repetition rates variable from 1 cps to 3 mc.



This instrument produces a pre-pulse and main pulse with delay variable from 90 milli-microseconds to 105 milli-seconds; Main pulse width is variable over the same range. Rise time is less than 10 millimicroseconds on the main pulse and a cable reflected pulse is available with a rise time of 8 milli-microseconds.

Aluminum Solder Brochure

Latest and most complete information available on the soldering of aluminum is presented in a new booklet by **Reynolds Metals Company,** Dept. PRD-6, Box 2346, Richmond 18, Va.

Newest addition to the Reynolds series of technical brochures, the handy-size reference includes data on soldering fluxes, irons and flames, and gives complete information on actual soldering methods such as hot plate, dip, furnace, friction, glass fiber brush and ultrasonic operations.

Types and properties of aluminum solders are explained, plus the corrosion of soldered joints and their performance in aluminum. Text is supported by diagrammatic illustrations.

Copies of "Soldering Aluminum" can be obtained by writing to the firm.

Power Supply Bulletin

A new 4 page bulletin PS-2009 recently released by **Deltron Inc.,** 2905 N. Leithgow St., Philadelphia 33, Pa., describes a series of 26 all solid state regulated dc power supplies primarily intended for built-in applications. Exact custom voltages are available on an off-the-shelf basis. Voltages from 0–100 vdc at currents from 0–20 amperes are covered by units in this series. The quality level of the instruments makes them suitable for computers, missile systems, communication equipment, strain gage applications and other similar uses.

(Continued on page 151A)

hese manufacturers have invited PROCEEDINGS eaders to write for literature and further technical nformation. Please mention your IRE affiliation.

(*Continued from page 146A*)

New VP's For Sensitive Research

Sensitive Research Instrument Corp., 310 Main St., New Rochelle, N. Y., announces the appointment of Marvin I. Steinberg and Leonard J. Patterson as executive vice presidents and members of the board of directors.

L. J. Patterson

M. I. Steinberg

Steinberg, sales manager, has been associated with the company for 14 years and will also continue in his present capacity until a successor has been appointed. He is a graduate of Cornell University.

Patterson, formerly special assistant to the Director of Engineering, was graduated from the University of Scranton, and has been with Sensitive Research for three years.

Capacitor For Missiles

Smallest molded mica capacitor ever produced, according to the manufacturer, is a device named the "Missilmite." It is 73 per cent smaller (size: 0.0077 cubic inch) and 69 per cent lighter (weight: $\frac{1}{2}$ gram) than the nearest comparable units, it was stated by engineers at **General Instrument's subsidiary, Micamold Electronics Manufacturing Corp.,** 65 Gouveneur St., Newark 4, N. J.

The component now "meets or exceeds" the requirements of MIL-C-5A and MIL-C-11272A and is available in characteristics "C," "D" and "E," it was stated.

Capacitance range is from 5 through 240 $\mu\mu$f. Tolerances range from ± 2 per cent to ± 20 per cent. Characteristics: Available in C or D characteristics in capacitance range of 5 through 50 $\mu\mu$f, in C, D or E characteristics in capacitance range from 51 through 240 $\mu\mu$f. Extended capacitance range to 350 $\mu\mu$f.

Working Voltage: at 85°C, working voltage is 500 vdc for these capacitance ranges, and at 125°C, it is 300 vdc. Extended capacitance range available at 50 vdc for operation at both 85°C and 125°C.

Weight is approximately $\frac{1}{2}$ gram. Size is 0.370 inch long, 0.190 inch wide and 0.110 inch thick.

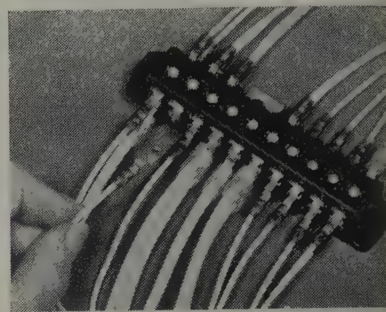
Terminal Block

Twin Lock Incorporated, 1024 W. Hillcrest Blvd., Inglewood, Calif., has announced the addition of a side entry terminal block to their line of electrical connectors.

(*Continued on page 152A*)

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 151A)

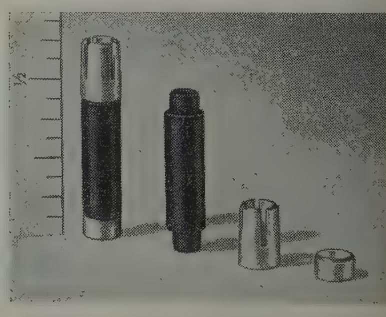


A proprietary item designated the T-1010, the new block retains all the inherent advantages featured in the original T-1000 vertical entry block. The difference in the T-1010 is the side entry feature which allows installation in flat, crowded spaces in which a vertical entry block cannot be used.

The block is designed for use with ground support equipment such as missile launchers, trailers and test stand equipment; however some are being specified for naval ordnance application. The new block is constructed of a molded phenolic base with reinforced barriers between the numbered terminal cavities; one cavity will accommodate four terminals. Up to 40 connections can be made with one block. In addition, adjacent terminals can be bussed together with Twin Lock jumpers. The block measures 5 inches long, by $1\frac{1}{4}$ inches wide, by 11/16 inches high. The terminal lugs (available in bulk or in strip bands of 1 to 3000 for automatic assembly equipment) and the block sockets are gold plated to meet environmental conditions of salt spray and humidity.

Coaxial Termination Resistors

This new product from **Filmohm Corp.,** 48 W. 25th St., New York 10, N. Y., is suited for use as a coaxial termination for $\frac{3}{8}$ line. It is said to be rugged, resistant to shock and vibration, and suitable for severe environmental conditions. Resistor design permits the use of metal contacts or fingers whose od is identical to the 0.125 inch od of the inner conductor of the line. Thus, the line inner conductor, metallic contact, and resistor, are all maintained at 0.125 inch od.



(Continued on page 154A)

NEWS New Products

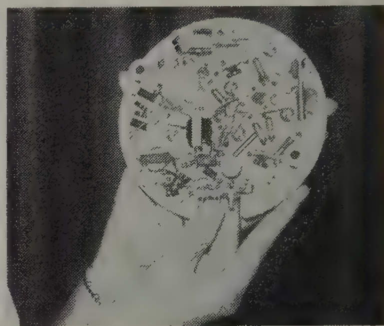
These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 152A)

When properly mounted in a $\frac{3}{8}$ line with type N connectors, a VSWR of 1.05 from dc to 4000 mc can be achieved. By the use of suitable impedance matching techniques the frequency range can be extended to 10,000 mc. Bulletin CT-1 offers detailed information. The resistors are now available from stock.

Telemetering Transmitter

An all-transistor 20 mc telemetering transmitter, designed for possible use in space satellites, has been built by **Philco Corp.**, C & Tioga, Philadelphia, Pa., for the U. S. Army.



(Continued on page 156A)

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 154A)

This is the first U. S. satellite transmitter to operate in the 20 mc band and it has the same type of modulation used by standard radio stations and amateur radio operators.

Developed by Philco's Research division for the Army Ordnance Missile Command, the satellite transmitter weighs less than 10 ounces, is five inches in diameter and one inch high.

New Philco high frequency power transistors were utilized in this telemetering transmitter which is said to provide up to one watt of output power. This is in contrast to one-to-two hundredths of a watt of output from transistorized transmitters in present U. S. satellites.

The transmitter consists of three sections:

Oscillator—Uses one high frequency power transistor in crystal controlled oscillator circuit.

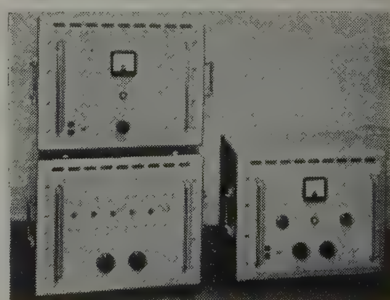
Power Amplifier—Uses four high frequency power transistors in parallel. Power gain of amplifier is 8 to 10 db.

Modulator—Uses one transistor as audio amplifier-driver and two transistors in Class B push-pull. Power gain of the modulator is in excess of 50 db. Modulation is applied to the collector of the rf amplifier.

Performance—Power output is about 600 milliwatts unmodulated carrier; 830 milliwatts with full modulation (nominal operating conditions). Transmitter can be adjusted for 1 watt output. Supply voltage —nominal 10.4 volts. Modulation frequency—400 cps to 2000 cps.

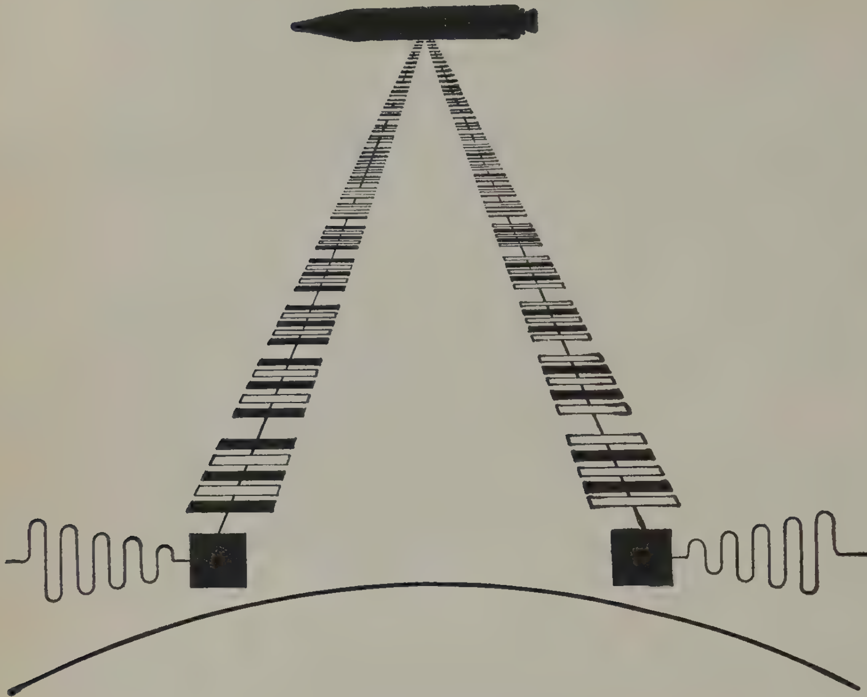
Noise Loading Test Set

Marconi Instruments, 111 Cedar Lane, Englewood, N. J., announce a new model of their Noise Loading Test Set Model 1249A. Three separate units, comprising



Noise Generator, Receiver and Filter Box enable slot and bandpass filters to be selected and plugged in at will. Built to CCIR specifications, the back to back ratio of the equipment is 70 db and the instrument can handle any number of channels up to 960.

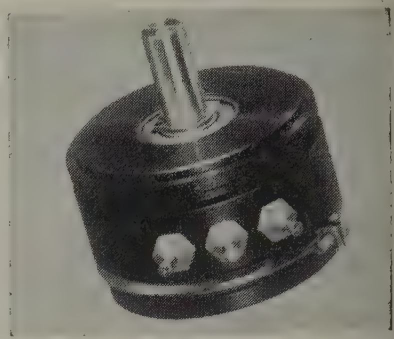
(Continued on page 158A)



(Continued from page 156A)

Precision Potentiometer

Helipot Div., Beckman Instruments, Inc., Fullerton, Calif., has just introduced a new all-metal $1\frac{7}{16}$ inch precision potentiometer with a power rating of 4.5 watts at 40°C, derated to 0 at 150°C.



Designated the Series 5400, these small, single-turn units are available in resistance ranges from 55 to 115K ohms. Standard linearity is ±0.5 per cent and tolerances as low as ±0.15 per cent can be supplied on special order. Non-linear versions, conforming to virtually any natural or empirical function, can be provided without external padding.

Series 5400 potentiometers are available with either sleeve or ball bearings, and shaft speeds up to 240 rpm will cause no damage. Ball bearing units have a standard torque of 0.6 ounce/inch, with 0.25 ounce/inch, available on special models.

All applicable sections of NAS 710, JAN-R-19, and MIL-E-5272A are met or exceeded. Up to 8 cups can be ganged on a single shaft, and as many as 12 taps can be added to each section.

Epoxy-Glass Laminates

A new, epoxy-glass laminate, designated INSUROK T-525 and T-525N, has been announced by **The Richardson Company,** 2700 Lake St., Melrose Park, Ill.

The new laminates meet military specification MIL-P-18177B, Type GEE, and NEMA G10.

Both grades are available copper-clad and non-clad. T-525 copper-clad is for normal printed circuit applications, while T-525N copper-clad will withstand cyanide solutions for gold plating, etc. Printed circuits on T-525 and T-525N can be made flush with adjoining surfaces by using heat and pressure. Standard sheet size is 36×42 inches and thicknesses $\frac{1}{32}$ inch to 1 inch for unclad INSUROK and $\frac{1}{32}$ inch to $\frac{1}{4}$ inch for copper-clad. Other thicknesses and sheet sizes are available on request.

Properties include excellent electrical characteristics at high humidity and exceptional dimensional stability. These INSUROK grades can be punched at room temperatures. Easily machined, they are drilled or tapped perpendicular or parallel to laminations with excellent results. Without special preparation printing or varnish is applied to the non-clad surface.

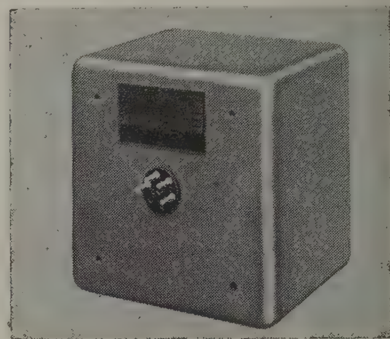
(Continued on page 160A)

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 153A)

Voltage Regulator

An ac line voltage regulator, especially designed for ground support and airborne equipment, has been introduced by **Kavamil Co., Inc.,** 1417 W. El Segundo Blvd., Compton, Calif. This unit, known as Model LVR-150, is one of a group of line voltage regulators being produced by the company which feature unusually wide frequency ranges. Electrical characteristics include input voltage: 90–135 vac; output voltage: 115 volt ±1 per cent for input of 90–135 volt; frequency range; 50–65 cps (also available for 900 cps); response time: approximately 100 milliseconds; ratings: 150 va (available from 50 va to 2.5 kva).

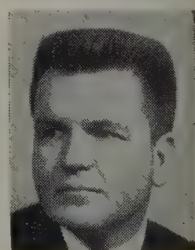


The Model LVR-150 is housed in a steel case and is completely encapsulated in epoxy for excellent vibration and shock characteristics. It is stud mounted with 4-pin connector. Size is 7×7×6 inches. Weight approximately 15 pounds. Available in lighter weights for airborne equipment.

Prices and detailed information may be had by writing to the firm.

O'Kane Ad Manager at Weinschel

Edward F. O'Kane has been named Advertising Manager for **Weinschel Engineering,** 10503 Metropolitan Ave., Kensington, Md., and for two new organizations which are closely associated with this rapidly growing manufacturer of precision microwave measurement equipment and electronic testing equipment. One of these new organizations, Flow Measurements Corp., is engaged in the development of thermal and ultrasonic flow meters. The other, The Kensington Scientific Education Society, Inc., will shortly introduce a series of educational kits in radio-electronics and other fields of science.



O'Kane has a wide background in both

(Continued on page 163A)

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NEWS New Products

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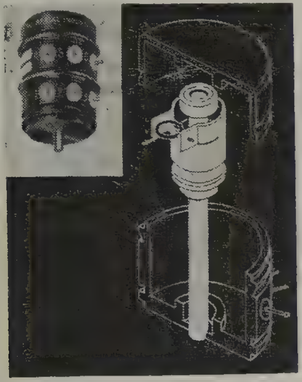
(Continued from page 160A)

...dustrial and consumer advertising. Prior ... joining Weinschel Engineering in Oc...ber 1958, he was advertising Manager of ...e Government Products Group of Ameri...n Machine & Foundry Company. From ...52 to 1955, Mr. O'Kane was with the ...emington Rand Division of Sperry Rand ...rporation, where he was responsible for ...e national advertising and sales promo...on program for the Remington portable ...pewriter. From 1945 to 1952, he was ...th Montgomery Ward & Company.

Precision Potentiometers

Continuous operation at extremely ...gh temperatures has been added to the ...erating characteristics of Dynamic Bal...ce single turn precision potentiometers ...eveloped and manufactured by **KIN...RONIC, div. Chicago Aerial Industries,** ...134 Pacific Ave., Franklin Park, Illinois. ... newly designed unitized contact arm en...neered for simplicity and efficiency en...bles reliable performance even at 225°C.



Already designed to withstand ex...remely severe conditions of shock, vibra...ion and acceleration, Kintronic potenti...meters incorporate a new contact arm ...ith all metal parts of stainless steel. The ...rm is dynamically balanced on the shaft ...nd contact assembly is dynamically bal...nced on arm. Dynamic Balance enables ...perating characteristics such as 5,000,000 ...ycle life, 2,000 cycle life at 30 g's, rota...ional speeds to 3,500 rpm and linear or ...unctional windings with 0.2 per cent max...num standard linearity and 0.1 per cent ...aximum standard linearity for larger ...izes.

For complete specifications contact ...intronic.

Blower

An extremely quiet dual centrifugal ...lower has been added to the line of pack...ged cooling equipment manufactured by **...cLean Engineering Laboratories,** Prince...on, N. J. The new blowers are designed ...pecifically for cooling computers, control ...anels, telemetry cabinets, consoles and ...ther equipment packed with tubes, power

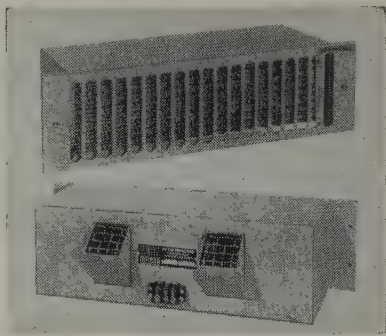
(Continued on page 164A)

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 163A)

supplies and heat generating equipment. The new units are designed to meet all important military specifications issued by the three services.



Air delivery is 150 to 500 cfm. The units fit 19 inch racks and the panel heights run from $3\frac{1}{2}$ inches to $8\frac{3}{4}$ inches in increments of $1\frac{3}{4}$ inches. The motors are capacitor types, available in 60 or 400 cps, single or three phase. The filters are a McLean permanent type. The case and grille conform to MIL-SPEC finishes. Data sheets on these and other packaged cooling units are available from the firm.

(Continued on page 168A)

ENGINEERS

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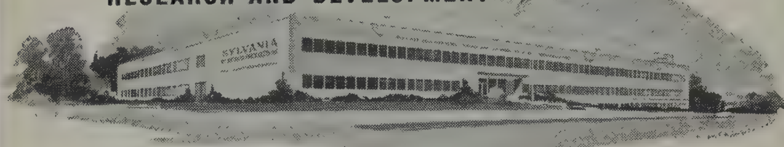

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SYSTEMS Operations research in systems problems, measures of effectiveness and optimal solutions; synthesis of block diagrams and formulation of mathematical descriptions; analysis by means of mathematical models of all aspects of complex systems, including the probablistic.

EQUIPMENT DEVELOPMENT From basic advanced communications systems concepts, working experimental equipment must be developed to operate effectively in feasible field test programs. Pulse techniques, transistor applications, digital data systems, radio frequency circuitry. Minimum of paper work.

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ENGINEERS

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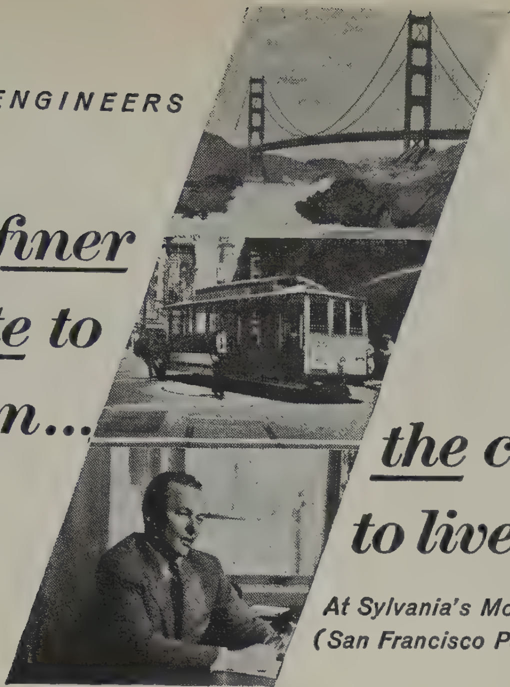

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P. O. Box 188—Mountain View, California

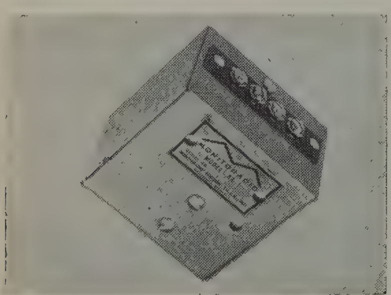
NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 164A)

Radiotelephone Alerting Device

A new inexpensive alert device for mobile dial direct radiotelephone systems has been developed by **Monitoradio Division, I.D.E.A., Inc.,** 7900 Pendleton Pike, Indianapolis 26, Indiana. Model AA-1 may be used to intermittently actuate any type of alarm where continuous operation of the alarm is not desirable. The alarm could be a light or bell or horn, etc.



For good reliability and economy the unit is powered by any receiver simply by hooking into receiver B+. While designed primarily for mobile applications Model AA-1 may also be used in any fixed location in conjunction with any receiver. Normal alarm cycle is 5 seconds off $\frac{3}{4}$ second on, but may be varied. In mobile dial direct or selective call systems the operator can be notified of call by connecting to the horn of the vehicle. Alarm cancels when call is acknowledged.

A modified version, when used with a regulated B power supply, can be used as a precision timer. In some applications it can also be used as an interrupter.

Model AA-1 is $2 \times 2\frac{7}{8}$ inches and weighs 12 ounces. Price is $19.95 user net.

Taylor Elected to Board

Trude C. Taylor, vice president, marketing, of **Telemeter Magnetics, Inc.,** was elected to the Board of Directors of the company at their January meeting. Taylor, a graduate of UCLA in engineering and of the Harvard Graduate School of Business, has been with the Telemeter Magnetics organization since 1957.

Taylor spent seven years prior to joining TMI with the Telecomputing Corporation where he served as vice president.

Telemeter Magnetics manufactures ferrite cores, core arrays, core storage buffers, computer memories, data translators and special purpose data handling systems.

New Bradley Plant

Bradley Semiconductor Corp. has moved into new headquarters, merging under one roof the production operations which were formerly divided between two separate plants in downtown New Haven, Conn. The general Bradley line of selenium and copper oxide rectifiers, modulators and arc suppressors has for a number of years been manufactured at the company's former main building, while the newly-introduced "Redtop" silicon diode has been coming from a nearby subsidiary plant.

The new Bradley building has been designed to the company's specific needs, and comprises 30,000 square feet on one level, ample for the company's present 160 production personnel. Tooled with the most modern equipment for developing, testing and producing electronic components in the semiconductor field, the new location also emphasizes efficiency in receiving and shipping. The $2\frac{1}{2}$-acre corner site allows for considerable future growth.

(Continued on page 170A)

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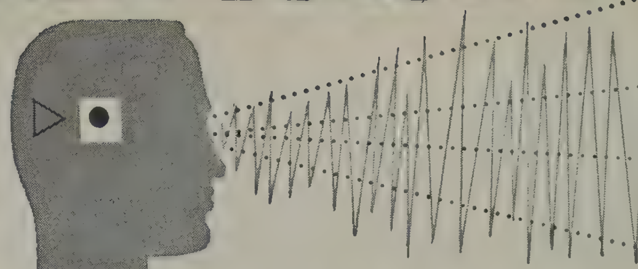

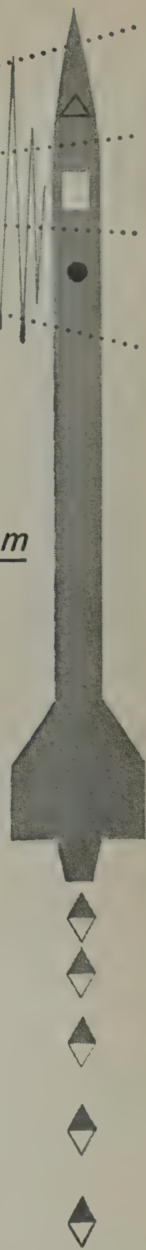

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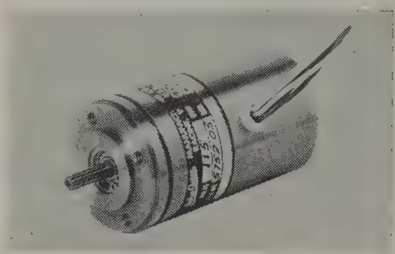
NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 168A)

Viscous Damped Motor



John Oster Manufacturing Co., Avionic Div., 1 Main St., Racine, Wis., has developed an adjustable viscous damped size 11 motor which is smaller, lighter and more economical than a motor tach used in feedback damping applications. Type 5752-05 consumes less power and presents no null or phasing problems in the feedback loop as compared with the motor tach which it replaces. Damping and gain may be independently adjusted. No load speed can be adjusted to any speed desired between 4800 and 7300 rpm depending upon damping characteristic required in the system. Unit can be built in a standard BuOrd Mark 14 characteristic. Meets MIL-E-5272A. Ambient temperature range −55°C to +125°C.

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(Continued on page 172A)

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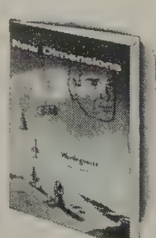

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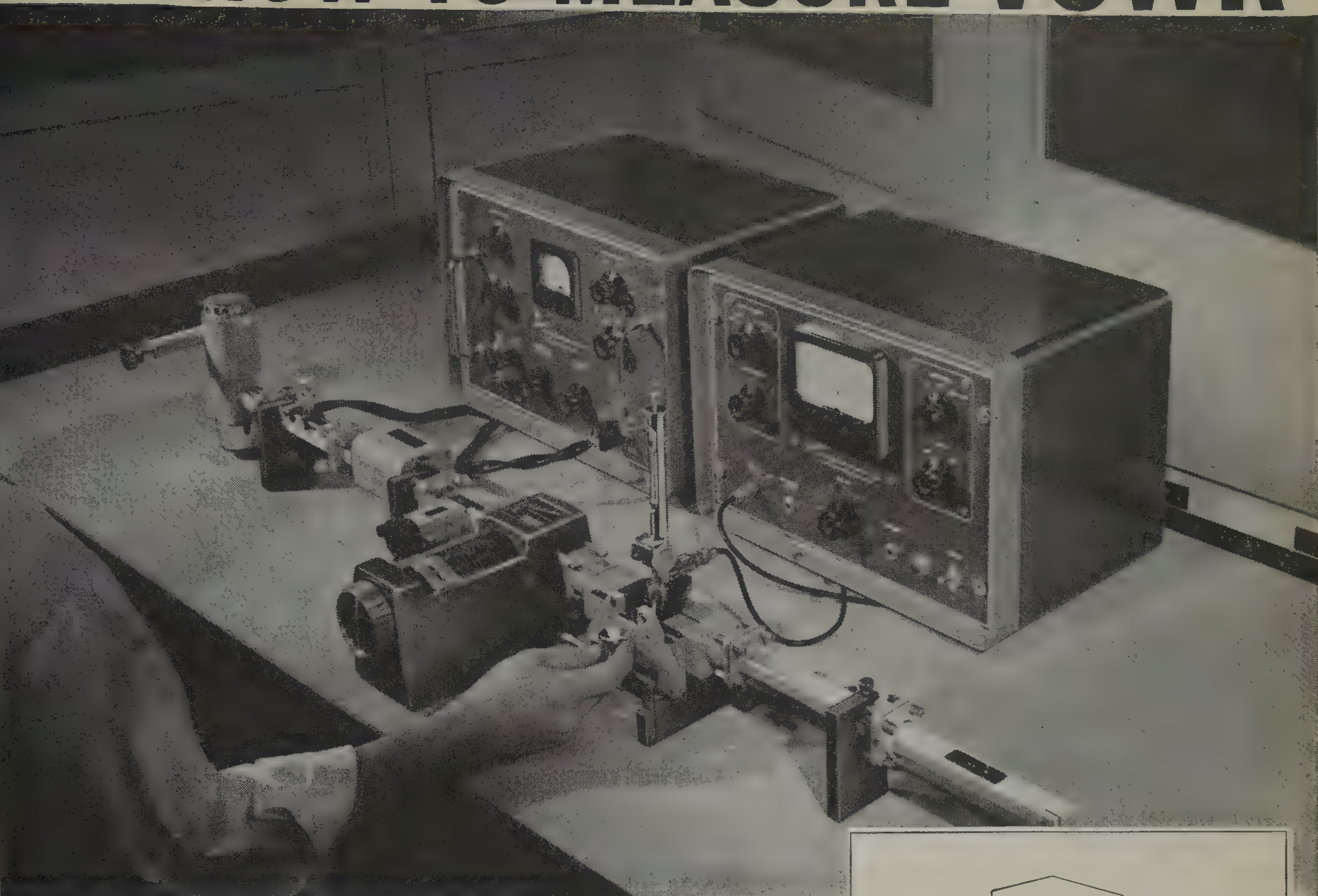
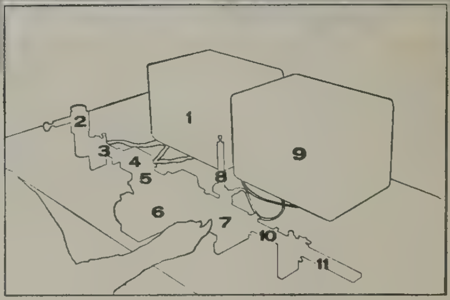
(*Continued on page 174A*)

HOW TO MEASURE VSWR





This microwave measurement bench is for the determination of Voltage Standing Wave Ratio using the slotted-line technique. Other systems utilizing directional couplers or magic tees for measurement of VSWR are known, but the use of the Slotted Section assures maximum accuracy.

Regardless of the technique used, accurate readings depend on the precision of the test instruments involved. When it comes to microwave test instruments PRD produces the widest range of the most precise equipment available anywhere in the world.

You will notice in the measurement bench shown that there are four test components separating the klystron tube mount from the Slotted Section. These are: A Slide Screw Tuner, ferrite Isolator, Level Set Attenuator, and a broadband direct reading Frequency Meter. THE USE OF THESE FOUR COMPONENTS IN THE TEST LINE IS MANDATORY FOR PRECISE VSWR MEASUREMENTS!

The reason for this is clear when you consider the interrelationship between VSWR, power, and frequency.

The Slide Screw Tuner is used to match the klystron output to that of the tandem test line, thereby maximizing its output and increasing its stability.

The use of the ferrite Isolator assures klystron frequency and power stability by shielding the source generator from changes in impedance further down the line. It accomplishes this with negligible attenuation of the incident power. The Level Set Attenuator is used to adjust the amount of power feeding the remainder of the test line.

The reaction Frequency Meter accurately monitors the output of the klystron by a resonant dip on the Standing Wave Amplifier.

A Slotted Section, tuned Broadband Probe, Standing Wave Amplifier, and matched Termination complete the precision waveguide, X-band, VSWR bench. A Klystron Power Supply to provide the signal source with power and modulation and a Fixed Waveguide Attenuator to simulate the unknown are also shown.

Special problems in VSWR and other related measurements?—Contact our Applications Engineering Department.

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For technical details and specifications covering products shown write:

TEST INSTRUMENTS USED IN THIS X-BAND VSWR BENCH

1—809 Klystron Power Supply, catalog page F-10
2—703 Shielded Tube Mount, catalog page F-8
3—303-A Slide Screw Tuner, catalog page B-14
4—1203 Isolator, catalog page A-21
5—159-A Level Set Attenuator, catalog page A-17
6—535 Frequency Meter, catalog page D-12
7—203-D Slotted Section, catalog page B-11
8—250-A Broadband Probe, catalog page B-12
9—277-A Standing Wave Amplifier, catalog page E-7
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(Continued on page 176A)

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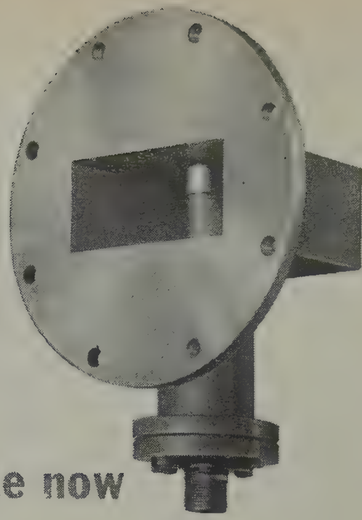

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Until recently signal simulators for monitoring radar receivers or microwave relays were of two types. One was a big and heavy ampere eater with cumbersome auxiliary equipment; and the other was a sensitive though delicate instrument suitable only for the laboratory.

We call your attention now to the Litton 2000 series of miniature gas noise sources. The Litton 2000 for waveguide use is pictured above. It has a first cousin, the Litton 2007 designed for coaxial cable use. We call your attention because most tubes in this series are now in production and we suspect there are frustrated design engineers who will receive this announcement with keen interest.

Our gas noise sources may properly be called miniature. They require only inches of space, smaller, lighter auxiliary equipment, and small voltages and currents. Around 500 volts fires them; 100 milliamperes maintains them. These characteristics, plus others, have caused them to find numerous applications: for in-flight calibration and test of aircraft microwave receivers; as *automatic* watchdogs on airborne radar systems; and in other systems which require various immunities to vibration, shock, humidity, and temperature cycling.

The Litton family of miniature gas noise sources, like all Electron Tube Division products, was designed to solve specific end item functions. We have found that this philosophy contributes to consistent reliability: tubes do their jobs more efficiently, for longer periods of time, and at lower overall cost to the buyer. Other advantages also result. For example, these noise sources require *no* ageing-in and the L-2000 is replaceable in the field without changing the mount.

Specific frequency ranges in L, S, C, X and K bands are covered. If you are concerned with radar transmission, or with microwave data links of any kind, we'll gladly send you more information. Write to Litton Industries Electron Tube Division, Office P5, 960 Industrial Road, San Carlos, Calif.

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(Continued on page 178A)

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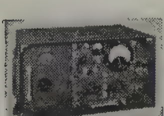

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(Continued on page 180A)

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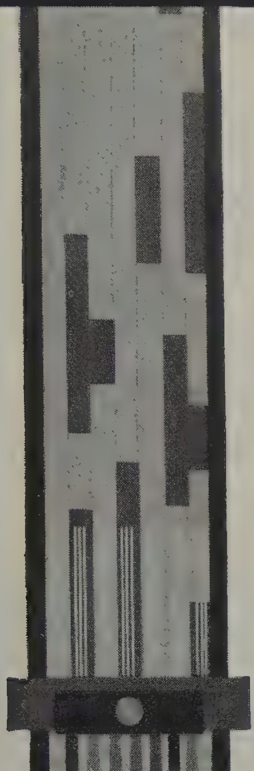

...Kearfott analog-to-digital converters

accuracy, reliability, varied capacities and codings

Kearfott's direct drive converters translate shaft rotation into electrical and visual digital form. These mechanically positioned units consist of coded drums, the number depending on the code and total count. The first, or units drum, connects directly to the input shaft and rotates as the shaft turns. On models where more than one drum is used, high speed odometer type gearing provides interconnection. All models can be read "on the run" or "on demand" and all tracks may produce simultaneous or serial readings, depending on the interrogating pulse.

The high performance and quality of Kearfott's analog-to-digital converters are assured by stringent testing for:

Mechanical Accuracy: The location of any code transition point must fall within one-third bit of its true angular position.

Low and High Temperature: All units operate efficiently at temperatures ranging from $-65°$ F to $+260°$ F.

Brush Skip: Conducted in conjunction with tests for temperature, vibration, shock and acceleration. A typical brush skip test employs 200 μsec., 23V DC amplitude interrogating pulses at 4000 cycles per second. With the shaft rotating, each output drum track is monitored for a total of 500,000 pulses, in which "lost" counts or "skips" will not exceed 1 in 15,000.

Shock: Kearfott converters are subjected to 18 shocks in three different planes with shock loading of $\pm$ 15 g's.

Vibration: Units must withstand vibrations between 5 and 500 cps with applied load of $\pm$ 10 g's.

Acceleration: All units are subjected to radial accelerations of varying magnitudes up to a maximum of $\pm$ 7.5 g's.

Life: Over 1 million linear feet (1000 hours at 82 RPM) of continuous brush travel on drum coded patterns can be achieved without excessive wear to drums, brushes or other sub-components.

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ANALOG-TO-DIGITAL CONVERTER CHARACTERISTICS

Kearfott Unit No.	423830-1	423607-1	Y1241-11A	P1240-11A	Y1240-11A	U1240-11	P1241-11A
Code (1)	B.D.	B.D.	B.D.	B.D.	B.D.	B.D.	C.B. (Gray)
No. of Drums	1	1	2	3	3	4	5
Range	0 to 84 in 348.5° of arc. 11.5° dead spot coded O	−19 to 0 to + 19	0 to 359	(+) 0 to (+) 999 (−) 999 to (−) 0	0 to 359.9	0 to 359.9	0-32,768 (2^{15})
Bits per Revolution	85	39	40	20	40	40	16
Revolutions for Total Range	1	1	9	100	90	90	2,048
Volts D.C.	23	12	23	23	23	23	10.5
Current (ma.)	20	20	20	20	20	20	20
Inertia (gm. cm.^{2}) (2)	158	48	104	28	158	100	20
Unit Diameter (in.)	2⅛	1¾	1$\frac{15}{16}$	1$\frac{7}{16}$	1$\frac{15}{16}$	1.875	1$\frac{7}{16}$
Unit Length (in.)	31/32	1$\frac{1}{16}$	1-43/64	1$\frac{15}{16}$	2$\frac{3}{16}$	2.8125	3
Life (3)	10^6 revolutions or 10^3 hours						
Static Torque (in.-oz.) (4)	0.5	0.1	1.0	.5	1.0	1.0	2 (break) 1 (running)
Weight (oz.)	5	3.5	5.75	4	7.75	6.5	5
Maximum Speed (RPM)	1100	250	300	400	300	700	600
Dielectric (Volts DC)	500	500	500	500	500	500	500

(1) B.D. (Binary Decimal), C.B. (Cyclic Binary).
(2) Inertia measured at maximum trip.
(3) Under recommended conditions.
(4) At room temperature.

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(Continued on page 186A)

Movable antenna at the Mullard Radio Astronomy Observatory, Cambridge University, England.

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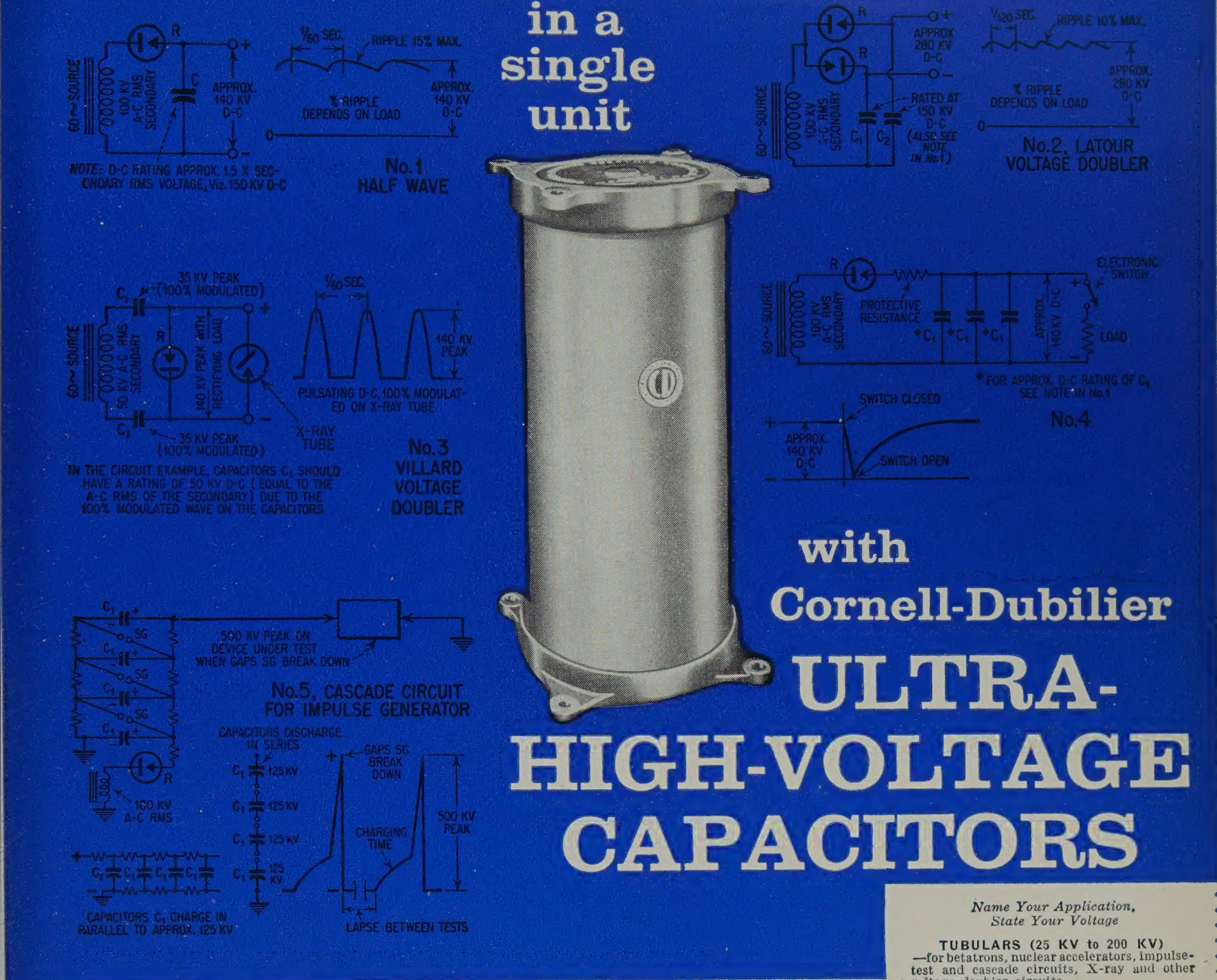

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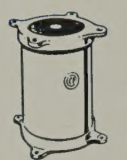

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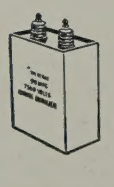

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